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BASIC ELECTRONICS

PREPARED BY

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PREFACE

Basic Electronics is written for men of the U. S. Navy and Naval Reserve whose duties require them to have a knowledge of the fundamentals of electronics. Electronics is defined as the science and technology that is concerned with devices involving the emission, behavior, and effect of electrons in vacuums, gases, and semiconductors. Technically speaking, electronics is a broad term extending into many fields of endeavor. Today, electronics projects itself into Navy life at every turn. It aims guns, drops bombs, navigates ships, and helps control engineering plants. It is therefore important to become well informed in basic electronics in order to be able to qualify for any of the many applicable rates or ratings.

As one of the NAVY TRAINING COURSES, this book was prepared by the U. S. Navy Training Publications Center for the Bureau of Naval Personnel.

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BASIC ELECTRONICS

CHAPTER 1

TUNED CIRCUITS

INTRODUCTION

Basic Electronics presents many of the basic concepts in the field of electronics. Emphasis is placed primarily on the theory of operation of typical electronic components and circuits that have frequent application in naval electronic equipments. The description of specific equipments is left to the rating texts.

This training course is intended as a basic reference for all enlisted personnel of the Navy whose duties require them to have a knowledge of the fundamentals of electronics. However, it is not intended that each rate of each rating concerned must study every chapter in the book. The chapters on amplifiers, for example, includes an introductory chapter which will be useful to all who are concerned with amplifiers, plus two other chapters which cover d-c amplifiers and audio-power amplifiers which will be necessary only for those who service and maintain those types of equipment. A suggested study guide indicating the ratings which should use the book, and the pertinent chapters for each follows:

Rating	*Chapters*
AQ	1, 2, 3, 4, 5, 6, 7, 8, 10, 11, 13, 14; appendix 2.
AT	ALL; appendix 2.
EM	2, 3; appendix 2.
ET	ALL; appendix 2.
FT	ALL; appendix 2.

GF	1, 2, 3, 4, 5, 6, 7, 8, 10, 11, 13; appendix 2.
GS	1, 2, 3, 4, 5, 6, 7, 8, 10, 11, 13; appendix 2.
IC	2, 3, 4, 5, 6, 13 (partial); appendix 2.
MN	2, 3, 4, 5, 7, 13 (partial); appendix 2.
RD	2, 3, 4, 7, 14; appendix 2.
RM	2, 3, 4, 7, 8, 9, 11 (partial), 12, 13; appendix 2.
SO	13; appendix 2.
TD	1, 2, 3, 4, 5, 6, 7, 8, 13; appandix 2.
TE	13; appendix 2.
TM	1, 2, 3, 4, 5, 6, 7, 8, 9, 12, 13; appendix 2.

In general, irrespective of the field, electrical networks comprise not more than four fundamental qualities—(1) resistance, (2) inductance, (3) capacitance, and (4) control devices such as vacuum tubes. *Basic Electronics* discusses the action of circuits and components in terms of these fundamental concepts and applies Ohm's law to the solution of related problems.

This training course introduces the subject of electronics as it is applied to tuned circuits and immediately involves the concept of resonance which is a basic quality of these circuits. Elementary mathematics, including algebra, geometry, and trigonometry is used to illustrate circuit behavior. Algebraic derivations are provided for those equations that require explanation. It should be understood, however, that the formula derivations are included only to strengthen the background of understanding of why particular components behave as they do under different circuit conditions. Those who find the formula derivations too difficult to follow should not be discouraged. If the reader acquires an understanding of WHAT the characteristics of particular circuits are and HOW the circuits behave, he is getting the main points. In other words, he can study most parts of this text with or without the mathematical derivations of formulas; he will acquire more understanding of the fundamentals of electronics if he uses both text and formulas. Throughout the text, emphasis is placed on circuit behavior.

A knowledge of the principles of basic electricity is espe-

cially important to an understanding of tuned circuits and the reader is urged to familiarize himself with these principles before attempting to read "Tuned Circuits." The principles of basic electronics include those of basic electricity and the transition between the two subjects is facilitated by selecting "Tuned Circuits" as the first chapter in *Basic Electronics*.

This training course next discusses the operating principles of vacuum tubes and their action in power supplies, amplifiers, and oscillators. The central portion of the text includes a discussion of modulation, detection, transmitters, receivers, antennas, and radio wave propagation. The text concludes with a discussion of electronic test equipment and an introductory chapter on radar. The student is urged to read this training course thoughtfully and deliberately with pencil and paper at hand, and to refrain from skimming the text or from overlooking the questions. It is hoped that the rhyme "This is the age of the half-read page" will not apply to those who read *Basic Electronics*.

INTRODUCTION TO TUNED CIRCUITS

A tuned circuit has capacitance, inductance, and resistance in series or in parallel. When the circuit is energized at a particular frequency, known as the resonant frequency, an interchange of energy occurs between the coil and capacitor. This interchange of energy tends to build up in amplitude far above the amount delivered by the energizing source. This is known as a resonant condition. At resonance, the inductor stores energy during the half cycle that the capacitor discharges and returns the energy during the next half cycle to recharge the capacitor. Because the circuit resistance acting in series with the inductor and capacitor is low, large amounts of energy may be exchanged at the resonant frequency, with minimum loss of energy in the circuit. The small energy loss incurred in the circuit is replenished from the source feeding the circuit.

At resonance, the time needed to charge the capacitor must be equal to the time needed to discharge the coil, otherwise

the charge and discharge will be out of step and cancellation will result.

In a series-tuned circuit, the impedance in ohms across the terminals of such a circuit is

$$Z=\sqrt{R^2+(X_L-X_C)^2}.$$

Since the inductive reactance, X_L, and the capacitive reactance, X_C, are equal and opposite in polarity at the resonant frequency, they balance each other and the actual total reactance is reduced to zero. Because $X_L-X_C=0$, the total impedance of the circuit at the resonant frequency is equal to the resistance of the circuit, or $Z=R$; at resonance, the maximum amount of current will flow in the circuit.

The most important characteristic of a SERIES-TUNED CIRCUIT is that AT RESONANCE THE CIRCUIT IMPEDANCE IS A MINIMUM.

At frequencies below resonance the series circuit acts like a capacitor plus a resistor, and accordingly the circuit current is reduced. At frequencies above resonance, the circuit acts like an inductor plus a resistor, and the current is likewise reduced.

At resonance, the voltage across the capacitor and the voltage across the inductor are equal in magnitude. Because they are 180° out of phase with each other their vector sum is zero and the source voltage appears across the circuit resistance. Because the series resistance is low, the source voltage is small in relation to the voltage across the coil and capacitor. Under these circumstances the voltage appearing across either the inductor or the capacitor may be much higher than the input voltage.

A parallel-tuned circuit consists of a combination of inductance and capacitance connected in parallel. A small value of resistance (representing the inherent resistance of the two components) may be considered as acting in series with the inductance and capacitance.

At resonance, the nonenergy component of the lagging current flowing through the inductive branch is exactly equal to the nonenergy component of the leading current

flowing through the capacitive branch. Under these circumstances the current flowing in either of the branches may be much greater than the line, or input, current. Because these currents are 180° out of phase, they neutralize. The small in-phase current that now flows in the line is due to the inherent resistance of the circuit. Therefore, at resonance the impedance offered by the parallel circuit is a MAXIMUM and is purely RESISTIVE.

The most important characteristic of a PARALLEL-TUNED CIRCUIT is that AT RESONANCE THE CIRCUIT IMPEDANCE IS A MAXIMUM.

At frequencies below resonance, the current through the inductive branch is large and lags the applied voltage by approximately 90°. At the same time a smaller component of current, which leads the applied voltage by approximately 90°, flows through the capacitive branch. Above resonance, the opposite conditions prevail.

If the ratio of the reactance to the inherent resistance of both the inductor and the capacitor is high, the circuit will be in resonance at the same frequency irrespective of whether the components are connected in series or in parallel.

In radio receivers, tuned circuits are used both for the selection of the desired frequency and for the rejection of undesired frequencies. The relative ability of a receiver to select the desired signal while rejecting all others is called SELECTIVITY.

In radio transmitters the entire process of radio-frequency power generation and amplification depends on the proper functioning of tuned circuits.

Test instruments such as signal generators, oscillators, and frequency meters, as well as other electronic devices such as television transmitters and receivers and radar and sonar equipments, employ many tuned circuits.

Before tuned circuits can be analyzed, an elementary understanding of vectors and vector algebra is required. Accordingly, a brief review of vectors as they are expressed both in the rectangular and the polar form follows.

EXPRESSING VECTORS ALGEBRAICALLY

Many common physical quantities such as temperature, the speed of a moving object, or the displacement of a ship can be expressed as a certain number of units. These units define only the magnitude and give no indication of the direction in which the quantity acts. Such quantities are called SCALAR quantities. If both the magnitude and the direction in which the quantity acts are indicated, it is called a VECTOR quantity and may be represented by a vector.

Electrical vectors are commonly used to represent a-c currents and voltages and their phase relations. The length of the vector represents the magnitude of the quantity involved and the direction of the vector, with respect to a reference axis, represents the lapse in time between the positive maximum values of current and voltage.

Impedance triangles, the sides of which represent vector quantities, are also used to represent the resistance and reactance components of a-c circuits. These are right triangles, having a base equal to the resistive component, an altitude equal to the reactive component, and a hypotenuse equal to the combined impedance. The angle between the combined impedance and the resistive component (hypotenuse and base) is equal to the phase angle between the voltage across the impedance and the current flowing through it.

In this chapter it is necessary to determine circuit impedances by the addition, subtraction, multiplication, and division of vector quantities. When it is inconvenient to express the quantity by simple algebra, a system of complex notation is used.

Imaginary Numbers

In calculations in electronics it is often necessary to perform operations involving the square root of a negative number—for example, $\sqrt{-9}$, $\sqrt{-5}$, and $\sqrt{-x}$. Because no number when multiplied by itself will produce a negative result, the roots of numbers such as the foregoing cannot be

extracted. It therefore becomes necessary to introduce a new type of notation to indicate the square root of a negative number. These numbers are called IMAGINARY NUMBERS to distinguish them from the so-called REAL NUMBERS. Actually, the numbers that we call imaginary in the mathematical sense are real in the physical sense. The term is merely one of convenience, as will be pointed out in the succeeding paragraphs.

In algebra, the foregoing quantities are treated as $\sqrt{-1}\sqrt{9}$, or $\sqrt{-1}\times 3$; $\sqrt{-1}\sqrt{5}$; and $\sqrt{-1}\sqrt{x}$. The term $\sqrt{-1}$ is expressed as i (for imaginary) in mathematics books, but when working with electrical circuits it is convenient to use the term j (called the J OPERATOR), because i is used to indicate the instantaneous value of the circuit current.

Graphical Representation

In order to present a quantity graphically, some system of coordinates must be employed. Quantities involving the j operator may be conveniently expressed by the use of RECTANGULAR COORDINATES, as shown in figure 1–1. In order to specify a vector in terms of its X and Y components, some means must be employed to distinguish between X-axis and Y-axis projections. Because the $+Y$-axis projection is $+90°$ from the $+X$-axis projection, a convenient operator is one that will, when applied to a vector, rotate it without altering the magnitude of the vector. Let $+j$ be such an operator that produces 90° COUNTERCLOCKWISE rotation of any vector to which it is applied as a multiplying factor. Also, let $-j$ be such an operator that produces 90° CLOCKWISE rotation of any vector to which it is applied as a multiplying factor.

Successive applications of the operator $+j$ to a vector will produce successive 90° steps of rotation of the vector in the counterclockwise direction without affecting the magnitude of the vector. Likewise, successive applications of the operator $-j$ will produce successive 90° steps of rotation in the clockwise direction. This rotation is shown in table 1.

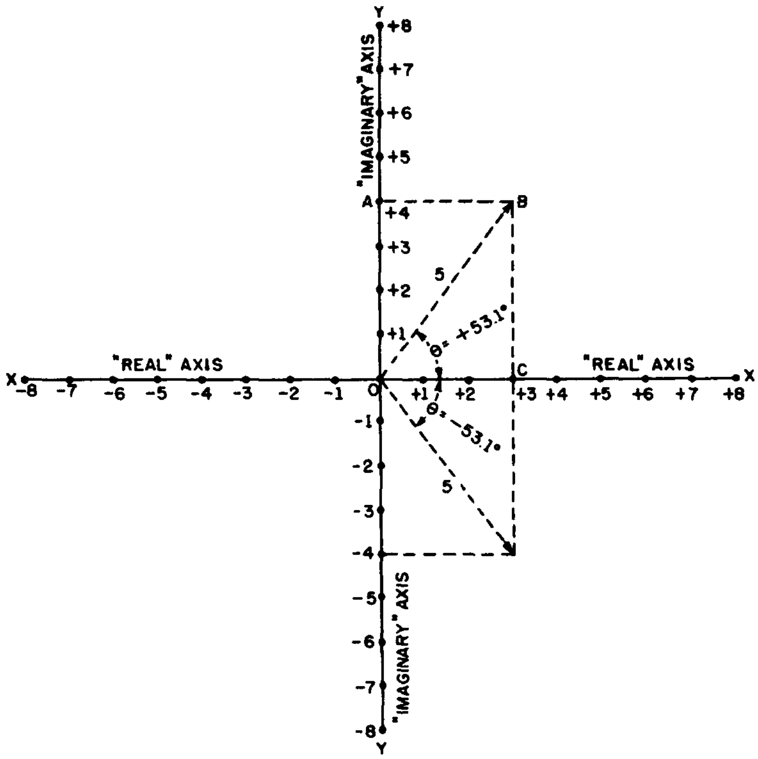


Figure 1-1.—Coordinates showing real and imaginary axes.

TABLE 1.—*Relation of operator* j *to vector rotation*

Operator	Mathematical equivalent	Direction of rotation	Degree of rotation
j	$\sqrt{-1}$	Counterclockwise	90
j^2	-1	do	180
j^3	$-\sqrt{-1}$	do	270
j^4	1	do	360
$-j$	$-\sqrt{-1}$	Clockwise	−90
$(-j)^2$	-1	do	−180
$(-j)^3$	$\sqrt{-1}$	do	−270
$(-j)^4$	1	do	−360

Consider the following example: The number, $+4$, in figure 1–1 indicates that 4 units are measured from the origin along the X axis in the positive direction. A $+j$ operator placed before the 4 indicates that the number is to be rotated 90° counterclockwise and will now be measured along the Y axis in a positive direction. Likewise, a $-j$ operator placed before the 4 indicates that the number is to be rotated 90° clockwise, and will now be measured along the Y axis in the negative direction.

It may be recalled that inductive reactance, X_L, is indicated as lying along the Y axis in the positive direction, and capacitive reactance, X_C, is indicated as lying along the Y axis in the negative direction; resistance in each case is measured along the X axis in the positive direction. Therefore, $+j$ has a direct association with X_L in that both are measured in the same direction along the Y axis, and $-j$ similarly has a direct association with X_C.

The function of the j operator may be shown as follows: The expression, 4 ohms, indicates that pure resistance is involved. In order to indicate that the 4 ohms represent capacitive reactance or inductive reactance a special symbol is needed. The use of the j operator gives a clear indication of the type of reactance. For example, if the j operator is not used, the 4 ohms is resistive. If $+j$ is used $(+j4)$, the 4 ohms is inductive reactance. If $-j$ is used $(-j4)$, the 4 ohms is capacitive reactance.

The so-called COMPLEX NUMBER contains the "real" and the "imaginary" terms connected by a plus or a minus sign. Thus, $3+j4$ and $3-j4$ are complex numbers. This means that the 3 and the 4 in each instance are to be added vectorially, and the $+j$ and $-j$ indicates the direction of rotation of the vector following it. The real number in these examples is 3 and could be represented by a line drawn three units out from the origin on the positive X (resistance) axis. The imaginary number, $+j4$, could likewise be represented by a line extended 4 units from the origin on the positive Y, or X_L, axis; and $-j4$ could be represented by a line extended 4 units from the origin on the negative Y, or

X_C, axis. The IMAGINARY, OR QUADRATURE, quantities (for example, the X_L and X_C values) are always assumed to be drawn along the Y axis, and the REAL quantities (for example, the R values) are always assumed to be drawn along the X axis.

Addition and Subtraction of Complex Numbers

Values that are at right angles to each other cannot be added or subtracted in the usual sense of the word. Their sum or difference can only be indicated, as is done in the case of binomials (an expression involving two terms). Thus, assume that it is desired to add $3+j4$ to $3-j4$.

$$\begin{array}{r} 3+j4 \\ \underline{3-j4} \\ 6-\ 0 \end{array}$$

The imaginary term disappears, and only the real term, 6, remains. If $3+j4$ is added to $3+j4$, the sum is the complex quantity, $6+j8$.

One complex expression may also be subtracted from another complex expression in the same manner that binomials are treated. For example, $3-j2$ may be subtracted from $3+j4$ as

$$\begin{array}{rr} & 3+j4 \\ (-) & \underline{3-j2} \\ & 0+j6 \end{array}$$

The real term disappears, and the result is 6 units measured upward from the origin on the Y axis. If $3-j2$ is subtracted from $6+j4$, the difference is the complex quantity, $3+j6$.

Multiplication and Division of Complex Numbers

Complex numbers are multiplied the same way that binomials are multiplied—for example, if $3-j2$ is multiplied by $6+j3$

$$
\begin{array}{r}
3-j2 \\
6+j3 \\
\hline
18-j12 \\
+j9-j^2 6 \\
\hline
18-j3-j^2 6
\end{array}
$$

Because $j^2=-1$, the product becomes $18-j3-(-1)6$, or $24-j3$.

Complex numbers may be divided in the same way that binomials containing a radical in the denominator are divided. The denominator is rationalized (multiplied by its conjugate—a term that is the same as the denominator except that it has the opposite algebraic sign before the j term), and the quotient is expressed as a term having only a real number as the divisor. For example, if $4+j3$ is divided by $2-j2$,

$$\frac{4+j3}{2-j2}=\frac{(4+j3)(2+j2)}{(2-j2)(2+j2)}=\frac{2+j14}{8}=\frac{1+j7}{4}=0.25+j1.75$$

Rectangular and Polar Forms

Sometimes it is more convenient to employ polar coordinates than rectangular coordinates. In **RECTANGULAR FORM** the vector is described in terms of the two sides of a right triangle, the hypotenuse of which is the vector. Thus, in figure 1-1 vector OB is described in rectangular form by the complex number $3+j4$. In other words, the end of the vector, OB, is 3 units along the $+X$ axis and 4 units along the $+Y$ axis, and its length is 5 units.

The vector, OB, may also be described if its length and the angle, θ, are given. When a vector is described by means of its magnitude and the angle it makes with the reference line it is expressed in the **POLAR FORM**. In this instance the length is 5 units and the angle, θ, is approximately 53°. The vector, OB, may then be expressed in the polar form as $5\angle+53°$. If the rectangular form is $3-j4$, the polar form is $5\angle-53°$.

The plus sign is shown with positive angles in this chapter in order to emphasize positive angles as contrasted with negative angles. The negative sign preceding the angle indicates clockwise rotation of the vector from the zero position.

Converting From One Form to the Other

Assume that the rectangular form is expressed by the complex number, $3+j4$. The angle, θ, and the actual length of the vector, OB, are not given. The length, OB, can be determined by use of the Pythagorean theorem ($OB=\sqrt{3^2+4^2}$), but it is usually simpler to determine first the angle, θ, by finding the angle whose tangent is $\frac{4}{3}=1.33$. The angle is 53° (approximately). The length of OB can then be readily determined as

$$OB=\frac{4}{\sin 53^\circ}=\frac{4}{0.798}=5,$$

and the vector may be expressed in the polar form as $5 \angle +53^\circ$.

If the vector is originally expressed in the polar form as $5 \angle +53^\circ$, it may be converted to the rectangular form by the use of cos 53° and sin 53°. In this instance the vector is 5 units in length and makes an angle of approximately 53° with the $+X$ axis. Thus,

$$\sin 53^\circ=\frac{BC}{5},$$

or

$$BC=5 \sin 53^\circ=5\times 0.799=4 \text{ (approx.)}; \cos 53^\circ=\frac{OC}{5},$$

or

$$OC=5 \cos 53^\circ=5\times 0.6=3.$$

Therefore with BC and OC known, the vector may be expressed as the complex number $3+j4$.

The polar form may be converted to the rectangular form more concisely in the following manner:

$$5 \angle +53° = 5 \cos 53° + j5 \sin 53°$$
$$= 5 \times 0.6 + j5 \times 0.8$$
$$= 3 + j4$$

Addition and Subtraction of Polar Vectors

Unless polar vectors are parallel to each other they cannot be added or subtracted algebraically. Therefore, the polar form is converted first to the rectangular form. Then the real components are added algebraically, and likewise, the imaginary components are added algebraically. Finally the result may be converted back to the polar form.

Multiplication and Division of Polar Vectors

The method of multiplying and dividing complex numbers by treating them as binomials and rationalizing the denominators may be simplified considerably by first converting the vectors into polar form and then proceeding to combine them in the following manner:

To obtain the product of two vectors, multiply the numbers representing the vectors in polar form and add their corresponding angles algebraically. The resultant vector is in polar form. Thus,

$$(5 \angle +53°)\ (5 \angle -53°) = 25 \angle 0°.$$

To obtain the quotient of two vectors, divide the numerator by the denominator as in ordinary division, then subtract algebraically the angle of the denominator from the angle in the numerator. The resultant vector is in polar form. Thus,

$$\frac{10 \angle +25°}{5 \angle -20°} = 2 \angle +45°.$$

SERIES RESONANCE

Series-Resonant Circuit

A series-resonant circuit is composed of a capacitor, an

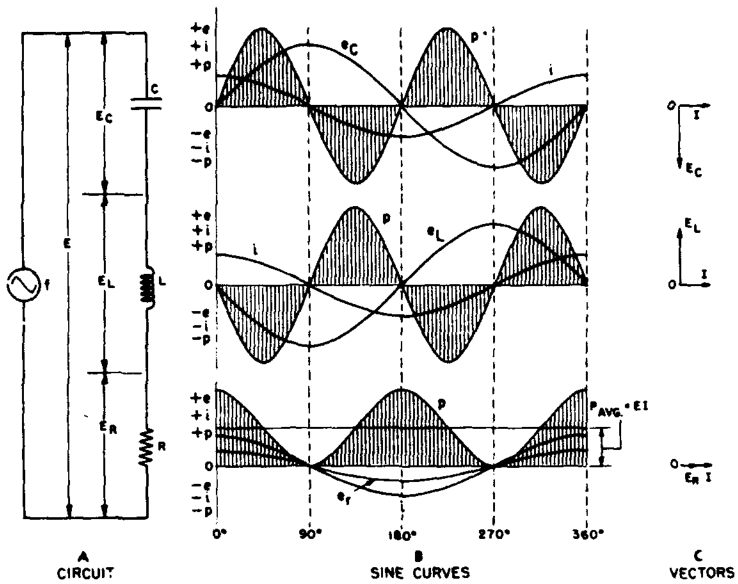


Figure 1-2.—Series resonance.

inductor, and a resistor, as shown in figure 1-2, A. The circuit losses that occur in the capacitor, the inductor, and the connecting leads are assumed to occur in the resistor, R, in the circuit.

Conditions Required for Series Resonance

In order for the series circuit shown in figure 1-2, A, to be in resonance, the frequency of the applied voltage must be such that $X_L = X_C$.

When a series circuit contains resistance, inductive reactance, and capacitance reactance, the total impedance for any frequency is given as

$$Z = R + j(X_L - X_C).$$

Because X_L increases and X_C decreases with an increase in frequency, at a certain frequency (the resonant frequency)

X_L will equal X_C, they will cancel, the j term will drop out, and Z will equal R. Therefore, at the resonant frequency, the power factor is unity. Furthermore, because the total impedance is now only the resistance, R, of the circuit, the circuit current is a maximum. In other words, at resonance the generator is looking into a pure resistance.

The resonant frequency of the series circuit is established as follows: At resonance

$$X_L = X_C \tag{1-1}$$

in which

$$X_L = 2\pi f L$$

and

$$X_C = \frac{1}{2\pi f C}.$$

Therefore, substituting the proper values in equation (1-1) gives

$$2\pi f L = \frac{1}{2\pi f C}. \tag{1-2}$$

Transposing (1-2)

$$f^2 = \frac{1}{4\pi^2 LC}; \tag{1-3}$$

and solving for f in (1-3),

$$f = \frac{1}{2\pi\sqrt{LC}},$$

where f is in cycles per second, L is in henrys, and C is in farads.

Where L is expressed in microhenrys (μh), C is expressed in micromicrofarads ($\mu\mu$f), and the frequency is expressed in megacycles (mc), a more convenient form of the equation is

$$f = \frac{159}{\sqrt{LC}}.$$

At frequencies below resonance, X_C is greater than X_L and the circuit contains resistance and capacitive reactance; at

frequencies above resonance, X_L is greater than X_C and the circuit contains resistance and inductive reactance. At resonance, the current is limited only by the relatively low value of resistance.

Because the circuit shown in figure 1-2, A, is a series circuit, the same current flows in all parts of the circuit, and therefore the voltage across the capacitor is equal to the voltage across the inductor, because X_L is equal to X_C. These voltages (fig. 1-2, C), however, are 180° out of phase, since the voltage across a capacitor lags the current through it by approximately 90° and the voltage across the inductor leads the current through it by approximately 90°. The total value of the input voltage, E, then appears across R and is shown as E_R in phase with the current, I.

Assume that at a given instant, corresponding to angle 0°, the current through the circuit is a maximum, as indicated in figure 1-2, B. During the first quarter cycle (from 0° to 90°) the circuit current falls from maximum to zero. The capacitor is receiving a charge, as is indicated by the rising voltage, e_c, across it. The product of the instantaneous values of e_c and i for this interval indicate a positive power curve. The shaded area under this curve represents the energy stored in the capacitor during the time it is receiving a charge.

During the first quarter of a cycle (0° to 90°), when the capacitor is receiving a charge, the magnetic field about the inductor is collapsing because the circuit current is falling and the inductor acts like a source of power that supplies the charging energy to the capacitor. The voltage, e_L, across the coil, is opposite in phase to the voltage building up across the capacitor and is shown below the line. Therefore, the product of the instantaneous values of the current and voltage across the inductor indicates a negative power curve for the coil between 0° and 90°.

During the second quarter cycle (90° to 180°) the capacitor discharges from maximum to zero, as indicated by the capacitor voltage curve, e_c, and the coil reverses its function and acts like a load on the capacitor. Thus, the capacitor now acts as a source of power. The product of a negative

current and a positive voltage (e_c) indicates a negative power curve for the capacitor for this interval. During the same quarter cycle the circuit is rising through the inductor (in the opposite direction) and energy is being stored in the magnetic field. The product of the negative current and negative voltage, e_L, for the second quarter cycle indicate a positive power curve for the inductor.

A similar interchange of energy between the capacitor and inductor takes place in the third and fourth quarter cycles. Therefore, the average power supplied to the inductor and capacitor by an external source is essentially zero. All circuit losses are assumed to be in the resistor, R. The voltage across the resistor and the current through it are in phase. The product of the voltage and current curves associated with the resistor indicate a power curve that has its axis displaced above the X axis. The displacement is proportional to the true average power which is equal to the product, EI (where E and I are effective values). Whatever power is dissipated in R is supplied by the source.

Quality, or Q

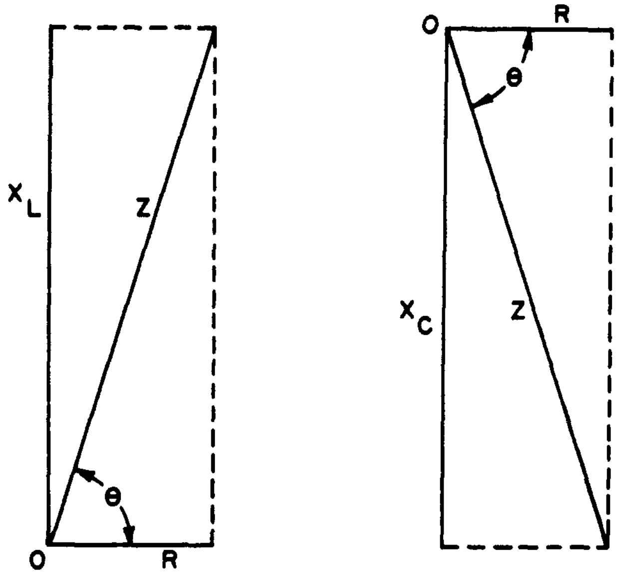
The ratio of the energy stored in an inductor during the time the magnetic field is being established, to the losses in the inductor during the same time is called the QUALITY, or Q, of the inductor; it is also called the FIGURE OF MERIT of the inductor. This ratio is

$$\frac{I^2X_Lt}{I^2Rt}=\frac{X_L}{R}=Q.$$

The Q of the inductor is therefore equal to the ratio of the inductive reactance to the effective resistance in series with it, and it approaches a high value as R approaches a low value. Thus, the more efficient the inductor, the lower the losses in it and the higher is the Q.

In terms of the impedance triangle (fig. 1–3, A)

$$Q=\frac{X_L}{R}=\tan\theta,$$



A

INDUCTIVE REACTANCE AND RESISTANCE

B

CAPACITIVE REACTANCE AND RESISTANCE

Figure 1-3.—Impedance triangles.

where θ is the phase angle between the hypotenuse, Z, and the base, R. As θ approaches 90°, tan θ approaches infinity, and the coil losses approach zero.

Similarly, in a capacitor the Q is a measure of the ratio of the energy stored to the energy dissipated in heat within the capacitor for equal intervals of time. This ratio is

$$\frac{I^2X_Ct}{I^2Rt}=\frac{X_C}{R}=Q,$$

where R is the effective resistance in series with the capacitive reactance, X_C (fig. 1-3, B). The effective resistance is low with respect to the capacitive reactance, and is such that when multiplied by the square of the effective capacitor

current equals the true power dissipated in heat within the capacitor.

Since most of the losses in a solid-dielectric capacitor occur within the dielectric rather than in the plates, the Q of low-dielectric-loss capacitors is high. The losses of an air-dielectric capacitor are negligible, and thus the Q of such a capacitor may have a very high value.

The Q of a CIRCUIT like the series-resonant circuit of figure 1–2, A, is the ratio of the energy stored to the energy lost in equal intervals of time. The expression becomes

$$Q=\frac{I^2X_Lt}{I^2Rt}=\frac{I^2X_Ct}{I^2Rt}=\frac{X_L}{R}=\frac{X_C}{R},$$

where R represents the total effective series resistance of the entire circuit. If the capacitor has negligible losses, the circuit Q becomes equivalent to the Q of the coil. The circuit Q may be maintained satisfactorily high by keeping the circuit resistance to a minimum.

INFLUENCE OF Q ON VOLTAGE GAIN.—In figure 1–2, A, the voltage across L is

$$E_L=IX_L=\frac{EX_L}{R}=QE.$$

The Q of the circuit is the ratio of the voltage across either the inductor or capacitor to that across the effective series resistance. In other words, the voltage gain of the series-resonant circuit depends on the circuit Q—that is

$$\text{V. G.}=\frac{E_L}{E}=\frac{IX_L}{IR}=\frac{X_L}{R}=\frac{IX_C}{IR}=\frac{X_C}{R}=Q.$$

REDUCTION IN VOLTAGE ACROSS C AND L NEAR RESONANCE.—Figure 1–4 shows the relation between the effective current and frequency in the vicinity of resonance for a series circuit containing a 159-μh coil, a 159-$\mu\mu$f capacitor. and an effective series resistance of either 10 ohms, or 20 ohms.

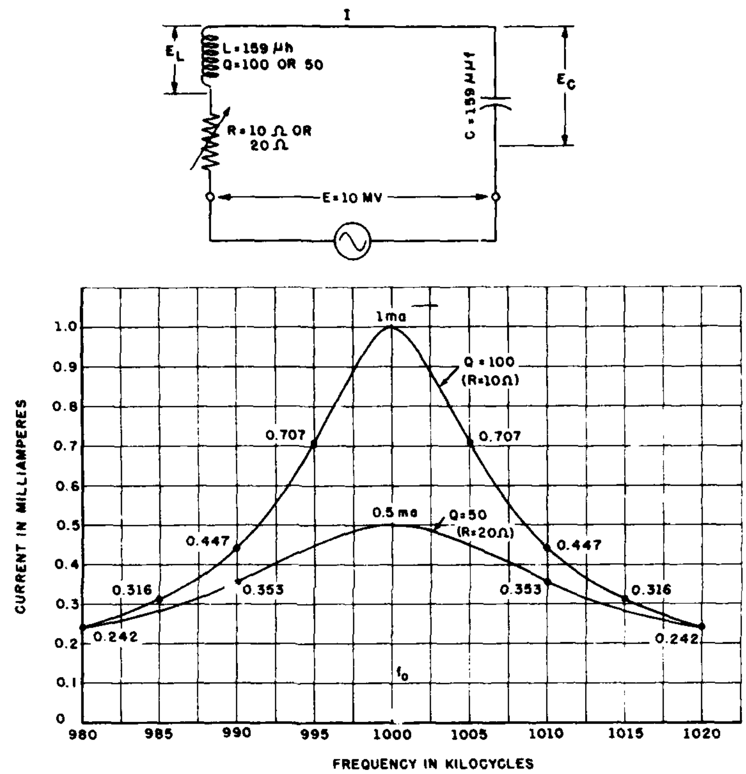


Figure 1–4.—Resonance curves of a series *L-C-R* circuit.

The resonant frequency, f_o, is

$$f_o=\frac{159}{\sqrt{LC}},$$

where f is in mc, L is in μh, and C is in μμf, or

$$f_o=\frac{159}{\sqrt{159\times159}}=1 \text{ mc.}$$

The reactances and impedance at resonance may likewise be determined. Thus,

$$X_{L_o}=2\pi fL=6.28\times10^6\times159\times10^{-6}=1{,}000\angle+90^\circ \text{ ohms};$$

$$X_{C_o}=\frac{1}{2\pi fC}=\frac{1}{6.28\times10^6\times159\times10^{-12}}=1{,}000\angle-90^\circ \text{ ohms}.$$

When $R=10$ ohms,

$$Z_o=R+jX_L-jX_C=10+j1{,}000-j1{,}000=10\angle0^\circ \text{ ohms}.$$

If the applied voltage is assumed to be 10 millivolts (mv) at a frequency of 1 mc, the circuit current is

$$I=\frac{E}{Z}=\frac{0.01}{10}=0.001, \text{ or } 1 \text{ ma}.$$

At the resonant frequency, the voltage across the inductor is

$$E_L=IX_L=0.001\times1{,}000=1, \text{ or } 1{,}000 \text{ mv},$$

and the voltage across the capacitor is the same, except it is 180° out of phase with the voltage across the coil. The losses in the coil and capacitor are assumed to be lumped in the effective series resistance. The circuit Q is

$$Q=\frac{X_L}{R}=\frac{1{,}000}{10}=100.$$

The voltage gain at resonance is

$$\frac{E_L}{E}=\frac{1{,}000}{10}=100.$$

If the frequency of the applied voltage is decreased by an amount $\frac{1}{2Q}$ times the resonant frequency, f_o, the current in the tuned circuit decreases to 0.707 of its value at the resonant frequency and leads the applied voltage by 45°. Thus, the input frequency is decreased an amount equal to

$$\frac{1}{2Q}\times f_o=\frac{1}{2\times100}\times1{,}000=5 \text{ kc},$$

and the new frequency is therefore 995 kc. At a frequency of 995 kc,

$$X_L=2\pi fL=6.28\times0.995\times10^6\times159\times10^{-6}=995\angle+90^\circ \text{ ohms},$$

and

$$X_C=\frac{1}{2\pi fC}=\frac{1}{6.28\times0.995\times10^6\times159\times10^{-12}}=1{,}005\angle-90^\circ \text{ ohms}.$$

The circuit impedance at 995 kc is

$$Z=10+j995-j1{,}005=10-j10=14.14\angle-45^\circ \text{ ohms}.$$

The circuit current at 995 kc is

$$I=\frac{E}{Z}=\frac{0.010}{14.14}=0.000707, \text{ or } 0.707 \text{ ma}.$$

At this frequency, the voltage across the coil, or the capacitor, is reduced to approximately 70 percent of its value at resonance because the *current* is reduced to this amount, and the reactance change is very small. The voltage across the coil is

$$E_L=IX_L=0.707\times995=705 \text{ mv}.$$

If the frequency of the applied voltage is decreased by an amount $\frac{1}{Q}$ times the resonant frequency, the current decreases to 0.447 of its value at resonance and leads the applied voltage by 63.4°. Thus, in the example of figure 1–4

$$\frac{1}{Q}\times f_o=\frac{1}{100}\times1{,}000=10 \text{ kc},$$

and 1,000—10=990 kc.

The inductive reactance at 990 kc is

$$X_L=2\pi fL=6.28\times0.990\times10^6\times159\times10^{-6}=990\angle+90^\circ \text{ ohms}.$$

and the capactive reactance is

$$X_C=\frac{1}{2\pi fC}=\frac{1}{6.28\times 0.990\times 10^6\times 159\times 10^{-12}}=1{,}010\angle -90^\circ \text{ ohms.}$$

The impedance of the series circuit at 990 kc is

$$Z=R+jX_L-jX_C=10+j990-j1{,}010=10-j20=22.4\angle -63.4^\circ \text{ ohms.}$$

At this frequency the circuit current is

$$I=\frac{E}{Z}=\frac{0.010}{22.4}=0.000447, \text{ or } 0.447 \text{ ma},$$

and the voltage across the coil is

$$E_L=IX_L=0.447\times 990=444 \text{ mv.}$$

Corresponding INCREASES in the frequency of the applied voltage above the resonant frequency will produce the same reductions in circuit current and voltage across the reactive portions of the circuit. In this case, however, the circuit current lags the applied voltage instead of leading it. Thus, the resonance curve is symmetrical about the resonant frequency in the vicinity of resonance.

If the circuit resistance is increased to 20 ohms, the Q is reduced to 50 and the resonance curve is flattened, as shown by the lower curve in figure 1–4. The series-resonant circuit amplifies the applied voltage at the resonant frequency. If the circuit losses are low the circuit Q will be high and the voltage amplification relatively large. For resonant circuits involving iron-core coils the Q may range from 20 to 100; for silver-plated resonant cavities at very high frequencies, the Q may range as high as 30,000. In practice, because nearly all of the resistance of a circuit is in the coil, the ratio of the inductive reactance to the resistance is especially important. The higher the Q of the coil, the better is the coil and the more effective is the series-resonant circuit that utilizes it.

If the circuit Q is low, the amplification at resonance is

relatively small and the circuit does not discriminate sharply between the resonant frequency and the frequencies on either side of resonance, as is shown by the lower curve in figure 1–4. The range of frequencies included between the two frequencies at which the current drops to 70 percent of its value at resonance is called the bandwidth for 70-percent response. A measure of the bandwidth for 70-percent response is $\frac{f_o}{Q}$. If the circuit Q is 100, the bandwidth for 70-percent response is

$$\frac{1{,}000}{100}=10 \text{ kc.}$$

Thus, if the frequency of the applied voltage is reduced from 1,000 kc to 995 kc or increased to 1,005 kc, the circuit current is reduced to 70 percent of its value at resonance. Likewise, the voltage across L or C is reduced to approximately 70 percent of its value at resonance. For the lower curve, representing a circuit having a Q of 50, the bandwidth is 20 kc.

Applications of Series-Resonant Circuits

Series-resonant circuits are used largely as filters (to be treated later in this chapter) for audio and radio frequencies. With proportionately larger component values the series circuit may be used as a power-supply filter. For example, assume that a d-c generator has a ripple frequency of 500 cps. A series-resonant circuit tuned to 500 cps may be connected across the terminals of the generator and thus effectively short-circuit the ripple voltage. The coil and capacitor insulation must be able to withstand the relatively high a-c voltages caused by the series-resonant action.

The series-tuned circuit may also be used to give an indication of frequency if the capacitor is calibrated for the appropriate frequency range. The capacitor and the inductor are connected in series with a current-indicating device across the source of the unknown frequency. At resonance the current, as indicated by the device, will be a maximum.

PARALLEL RESONANCE

Parallel-Resonant Circuit

A parallel-resonant circuit consists of a combination of inductance, resistance, and capacitance in two parallel branches as indicated in figure 1–5, A. Because the losses

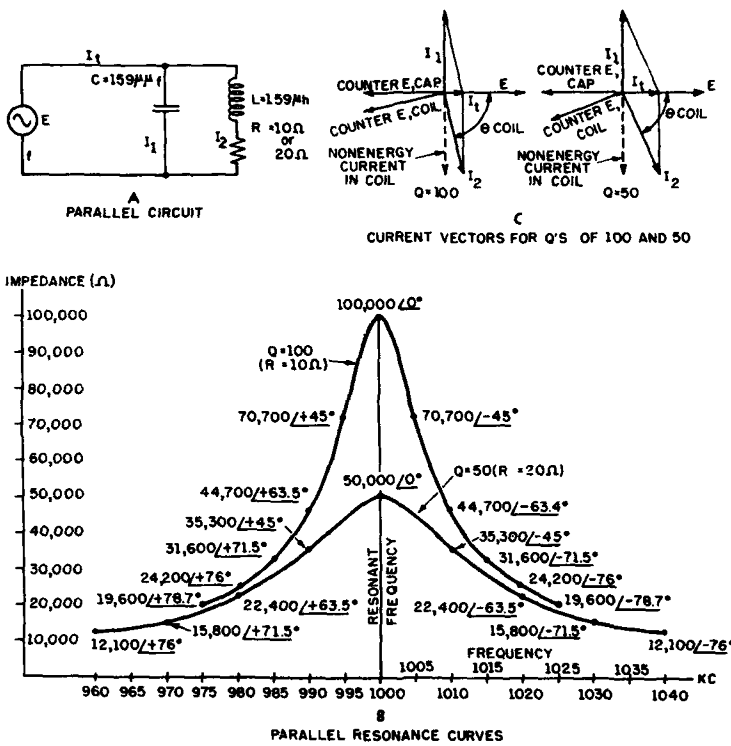


Figure 1–5.—Parallel resonance.

of the circuit are generally associated with the inductor, this branch includes a series resistor, *R*, in which all the losses are lumped. The other branch consists of a capacitor having negligible loss. At resonance, the same interchange of energy occurs between the capacitor and the inductor that occurs in the series-resonant circuit. The circuit impedance

vs. frequency is shown in figure 1–5, B, and the vector diagram is shown in figure 1–5, C.

Conditions Required for Parallel Resonance

At resonance the current, I_1, in the capacitive branch equals the nonenergy (reactive) component of current in the inductive branch as shown in the vector diagram of figure 1–5, C. Because the capacitor current leads the applied voltage by 90° and the inductor reactive component of current lags the applied voltage, E, by 90°, their vector sum is zero; and the line current, I_t, represents only the relatively small value of energy current that flows in the inductive branch. Thus, the parallel-resonant circuit has a high input impedance, and the line current is in phase with the applied voltage, which is the condition of unity power factor.

Formulas for f_o and Z_o

For all practical purposes the resonant frequency of a parallel circuit having a Q of 10 or higher may be expressed as

$$f_o=\frac{1}{2\pi\sqrt{LC}},$$

where f_o is in cycles, L is in henrys, and C is in farads. This is the same as the formula for the series-resonant circuit, because here too there is an approximate equality between the inductive and capacitive reactances.

The combined impedance, Z_o, at resonance for the two-branch parallel circuit is

$$Z_o=\frac{E}{I_t}=\frac{E}{I_2\cos\theta}=\frac{E}{\frac{E}{Z_2}\cos\theta}$$

$$=\frac{E}{\frac{E}{Z_2}\cdot\frac{R}{Z_2}}=\frac{Z_2^2}{R}. \qquad (1\text{–}4)$$

where Z_2 is the impedance of the L–R branch.

At resonance the capacitor current, I_1, is equal to the nonenergy component of the inductor current, I_2. Thus,

$$I_1 = I_2 \sin \theta$$

$$\frac{E}{X_C} = \frac{E}{Z_2} \cdot \frac{X_L}{Z_2}$$

$$Z_2^2 = X_C X_L. \tag{1-5}$$

Substituting (1–5) in (1–4),

$$Z_o = \frac{X_L X_C}{R} = \frac{\omega L \frac{1}{\omega C}}{R} = \frac{L}{CR},$$

where $\omega = 2\pi f$. Thus, the $\frac{L}{C}$ ratio (R being constant) is a factor which determines the magnitude of the impedance of the parallel circuit at the resonant frequency. The impedance-frequency curves (figure 1–5, B) for a parallel circuit have the same shape as the current-frequency curves for a series circuit. Note that the impedance across the terminals of the parallel circuit is maximum at resonance, whereas it is minimum for the series circuit. As in the series circuit, the resonance curves are sharper when the internal resistances are smaller, and the Q's are higher.

For example, the two-branch parallel-tuned circuit, figure 1–5, A, has a capacitor of 159 $\mu\mu f$ with negligible losses in parallel with a 159-μh coil having an effective resistance of 10 ohms. The resonant frequency is

$$f_o = \frac{159}{\sqrt{LC}} = \frac{159}{\sqrt{159 \times 159}} = 1 \text{ megacycle.}$$

The impedance at resonance is

$$Z_o = \frac{L}{CR_1} = \frac{159 \times 10^{-6}}{159 \times 10^{-12} \times 10} = 100{,}000 \text{ ohms.}$$

The coil reactance at the resonant frequency is

$$X_L = 2\pi fL = 6.28\times10^6\times159\times10^{-6} = 1{,}000 \text{ ohms}.$$

The coil Q is

$$\frac{X_L}{R_1} = \frac{1{,}000}{10} = 100.$$

Thus, as the parallel circuit strikes resonance, there is a rise in the combined impedance equal to Q times the coil reactance. The lower the coil resistance, the higher will be the coil Q and the combined impedance at the resonant frequency. The vector diagram for this example is shown at the left in figure 1–5, C.

Parallel Impedance Near Resonance

The parallel impedance, Z_t, of a parallel circuit at any frequency is

$$Z_t = \frac{Z_1 Z_2}{Z_1 + Z_2},$$

Where Z_1 is the impedance of one branch and Z_2 is the impedance of the other branch.

Below resonance—for example, at 995 kc—the impedance of branch 1 is

$$\frac{1}{\omega C} = \frac{1}{6.28\times995\times10^3\times159\times10^{-12}}$$

$$= 1{,}005\angle -90^\circ,$$

and the impedance of branch 2 is

$$R + j\omega L = 10 + j6.28\times995\times10^3\times159\times10^{-6}$$

$$= 10 + j995, \text{ or } 995\angle +90^\circ.$$

The parallel impedance is

$$Z_t = \frac{(1{,}005\angle -90^\circ)(995\angle +90^\circ)}{0 - j1{,}005 + 10 + j995}$$

$$=\frac{1{,}000{,}000\angle 0^\circ}{14.14\angle -45^\circ}$$

$$=70{,}700\angle +45^\circ.$$

In this example, the frequency deviation is $\frac{f_o}{2Q}$, or $\frac{1{,}000}{2\times 100}=5$ kc. When the frequency is deviated, an amount equal to $\frac{f_o}{2Q}$, the impedance falls to 70 percent of its value at resonance and the phase angle increases from 0° to 45°. Above resonance, the angle is negative; below resonance it is positive. In other words, above the resonant frequency, the circuit acts like capacitance in series with resistance; below resonance it acts like inductance in series with resistance.

Loading the Parallel-Resonant Circuit

Increasing the resistance (from 10 to 20 ohms) in series with the coil lowers the coil Q from 100 to 50. Since there are negligible losses in the capacitor the circuit Q is halved and the total impedance of the parallel circuit at the resonant frequency is

$$Z_o=\frac{L}{CR_2}=\frac{159\times 10^{-6}}{159\times 10^{-12}\times 20}$$

$$=\frac{1{,}000{,}000}{20}=50{,}000 \text{ ohms}.$$

Thus, the parallel impedance at resonance varies inversely with the resistance in the coil branch. This series resistance may represent the load on the parallel circuit. Hence, an increase in series resistance in the coil circuit may represent an increased load and a decrease in the total impedance of the parallel circuit. The vector diagram for this example is shown at the right in figure 1–5, C.

The Q of a parallel-tuned circuit at resonance may be

defined as the ratio of either the current in the capacitive or inductive branch, $I_{1\text{-}2}$, to the line current, I_t, or

$$Q=\frac{I_{1\text{-}2}}{I_t}.$$

When a load is inductively coupled to the tuned circuit, the load, in effect, adds resistance within the tuned circuit. The impedance of the circuit is thereby reduced, and the line current is increased, thus lowering the circuit Q. If a load is connected in shunt with the tuned circuit, the line current is increased and the circuit Q is lowered.

As long as the circuit Q is maintained above 10, the increased load which the increased series resistance represents does not materially affect the phase angle between line current and line voltage and does not change the resonant frequency. The increase in line current, as shown at the right in figure 1–5, C, as a result of the increase in resistance of the coil branch from 10 ohms to 20 ohms, is the result of the slight decrease in phase angle θ for the coil and the corresponding slight shift in the phase angle between the counter emf of the coil and the applied voltage.

The parallel-resonant circuit is often called a TANK CIRCUIT because it acts like a storage tank when used in some electron-tube circuits. The inertia effect of the inductor gives the tank a FLYWHEEL effect that permits the alternating current to build up in the tank. The relatively large current in the tank is equal to the circuit Q times the line current; the amplification of current is like the gain in momentum of a flywheel as it is being accelerated. The high input impedance of the parallel-resonant circuit is the result of the relatively large inductive emf of the inductor and the capacitive emf of the capacitor, both in approximate phase opposition with respect to the source voltage (fig. 1–5, C).

The parallel circuit is frequency sensitive to a varying degree, depending on the Q of the circuit. Below resonance, the lower impedance of the inductive branch causes the line current to increase and to lag the applied voltage. Con-

versely, above resonance the lowered impedance of the capacitive branch causes the line current to again increase and to lead the applied voltage. At resonance, the impedance is high and resistive. Generally the tank circuit is supplied by a relatively high-impedance source compared with series-resonant circuit sources. As the frequency of the source voltage is varied from below to above the resonant frequency, the voltage rises across the tank at the resonant frequency, and the line current falls as the tank current rises. The rise in voltage across the tank as resonance is approached is due to the decrease in line current and internal voltage loss of the source.

Frequently the tank circuit is used to couple energy into a load by utilizing the inductor as the primary of a transformer with the secondary connected to the load. When the secondary is tuned to the resonant frequency of the tank, the secondary current becomes a maximum. The field of the secondary current cuts the primary inductor and induces a counter emf in that coil. As mentioned previously, this action is equivalent to adding effective resistance in series with the inductive branch. Thus, coupling a load to the tank through mutual inductance lowers the parallel impedance and increases the line current. Coupling the load in this manner tends to slightly detune the tank so that it is generally necessary to retune by adjusting the capacitor. Loading the tank lowers the circuit Q, the parallel impedance, the tank current, and the voltage across the tank. At the same time the line current is increased.

Applications of Parallel-Resonant Circuits

The parallel-resonant circuit is one of the most important circuits used in electronic transmitters, receivers, and frequency-measuring equipment.

The i-f transformers of radio and television receivers employ parallel-tuned circuits. Parallel-tuned circuits are also used in the driver and power stages of transmitters, as well as in the oscillator stages of transmitters, receivers, and frequency-measuring equipment.

Various types of filter circuits employ parallel-tuned circuits as well as series-tuned circuits.

TUNED CIRCUITS AS FILTERS

Tuned circuits are employed as filters for the passage or reject ion of specific frequencies. Band-pass filters and band-rejection filters are examples of this type. Tuned circuits have certain characteristics that make them ideal for certain types of filters, especially where high selectivity is desired. A series-tuned circuit offers a low impedance to currents of the particular frequency to which it is tuned and a relatively high impedance to currents of all other frequencies. A parallel-tuned circuit, on the other hand, offers a very high impedance to currents of its natural, or resonant, frequency and a relatively low impedance to others.

Band-Pass Filters

A band-pass filter is designed to pass currents of frequencies within a continuous band, limited by an upper and lower cutoff frequency, and substantially to reduce, or attenuate, all frequencies above and below that band. A simple band-pass filter is shown in figure 1–6, A.

The series- and parallel-tuned circuits are tuned to the center frequency of the band to be passed by the filter. The parallel-tuned circuit offers a high impedance to the frequencies within this band, while the series-tuned circuit offers very little impedance. Thus, the desired frequencies within the band will travel on to the load without being affected; but the currents of unwanted frequencies—that is, frequencies outside the desired band—will meet with a high series impedance and a low shunt impedance so that they are in a greatly attenuated form at the load.

There are many circuit arrangements for both band-pass and band-elimination filters. However, for the purpose of a brief analysis the band-pass circuit shown in figure 1–6 will be considered. Let it be assumed that a band of frequencies extending from 90 kc to 100 kc is to be passed by the filter.

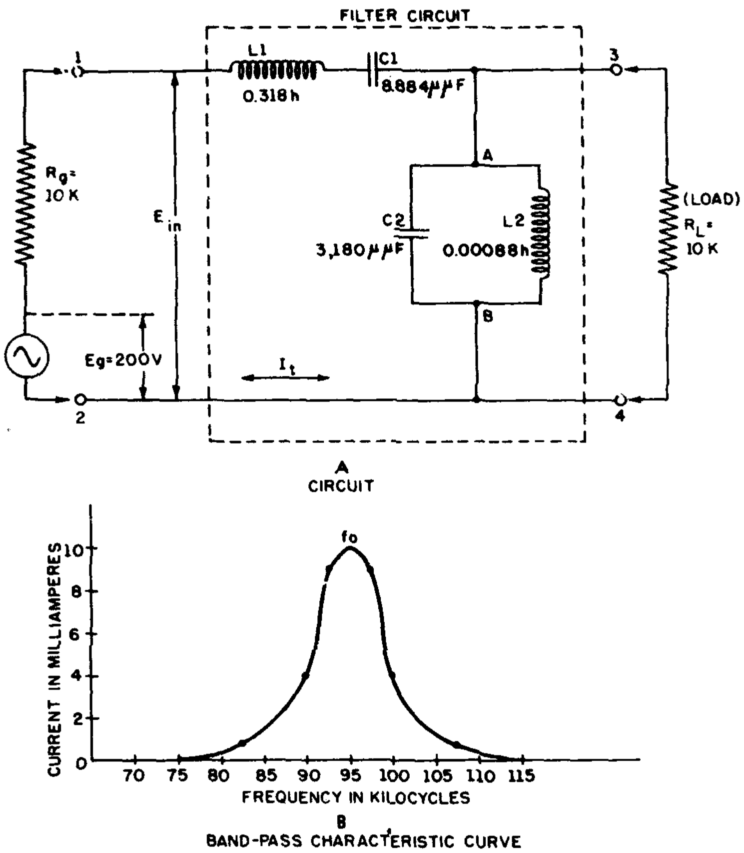


Figure 1-6.—Band-pass filter.

For an input and output resistance of 10,000 ohms, the values of inductance and capacitance are as indicated in the figure. The formulas by which these values are obtained may be found in handbooks on the subject.

The resonant frequency of the series circuit, $L1C1$, is

$$f_o = \frac{1}{2\pi\sqrt{LC}}$$

$$f_o = \frac{1}{6.28\sqrt{0.318 \times 8.884 \times 10^{-12}}}$$

$$= \frac{0.159 \times 10^{-6}}{\sqrt{2.81}} = \frac{159{,}000}{1.68} = 95{,}000 \text{ cps},$$

and for the parallel circuit, $L2C2$, is

$$f_o = \frac{0.159 \times 10^{-6}}{\sqrt{3{,}180 \times 10^{-12} \times 8.8 \times 10^{-4}}} = 95{,}000 \text{ cps}.$$

Thus both circuits are resonant at the center frequency of the band-pass filter; the upper limit of which is 100 kc and the lower limit 90 kc.

At resonance the impedances of $L1$ and $C1$ cancel and maximum current flows through the load, R_L; also, the parallel circuit, $C2L2$, offers almost infinite impedance and may be considered an open circuit. The inherent resistances associated with the filter components are neglected. Thus, at resonance, with an assumed source voltage of 200 volts and a total impedance of 20,000 ohms, the current through the load is approximately

$$I = \frac{E}{R} = \frac{200}{20{,}000} = 0.01 \text{ a, or } 10 \text{ ma}.$$

Below resonance—for example, at 90 kc—the impedances of $L1C1$ and $C2L2$ are such that with the assumed source voltage of 200 volts only about 4 milliamperes flow through the load. Further below resonance the current through the load is even less; and at 75 kc the load current drops to approximately 0.0024 milliamperes.

The same relative decrease in current occurs through the load with a corresponding increase in frequency. Figure 1–6, B, is a graph of the load current vs frequency characteristic of the filter shown in figure 1–6, A.

The current through the 10 k-ohm load (fig. 1–6, A) for an

arbitrarily chosen frequency of 92.5 kc may be determined as follows:

$$I_L = \frac{E_{3,4}}{R_L}, \tag{1-6}$$

where I_L is the current through the load, $E_{3,4}$ is the voltage across the load, and R_L is the load resistance.

$$E_{3,4} = I_t Z_{3,4}, \tag{1-7}$$

where I_t is the total current supplied by the generator and $Z_{3,4}$ is the combined impedance of the parallel circuit, $C2L2$, and the load, R_L.

$$I_t = \frac{E_g}{Z_t}, \tag{1-8}$$

where E_g is the voltage of the generator, and Z_t is the total impedance of the circuit.

Z_t may be determined by solving for the parallel impedance of $C2$ and $L2$, combining it with R_L, and then combining this impedance ($Z_{3,4}$) with the impedance of $C1$, $L1$, and R_g. Thus,

$$Z_t = \frac{\dfrac{(X_{L2}\angle{+90^\circ})(X_{C2}\angle{-90^\circ})}{+jX_{L2}-jX_{C2}}(R_L\angle 0^\circ)}{\dfrac{(X_{L2}\angle{+90^\circ})(X_{C2}\angle{-90^\circ})}{+jX_{L2}-jX_{C2}}+R_L} + jX_{L1} - jX_{C1} + R_g$$

$$= \frac{\dfrac{(510\angle{+90^\circ})(541\angle{-90^\circ})}{+j510-j541}\times 10{,}000\angle 0^\circ}{\dfrac{(510\angle{+90^\circ})(541\angle{-90^\circ})}{j510-j541}+10{,}000} + j185{,}000 - j194{,}000 + 10{,}000$$

$$= \frac{(8{,}900\angle{+90^\circ})(10{,}000\angle 0^\circ)}{10{,}000+j8{,}900} + 10{,}000 - j9{,}000$$

$$= \frac{(8{,}900\angle{+90^\circ})(10{,}000\angle 0^\circ)}{13{,}370\angle{+41.7^\circ}} + 10{,}000 - j9{,}000$$

$$= 6{,}670\angle{+48.3^\circ} + 10{,}000 - j9{,}000$$

$$=4{,}450+j4{,}980+10{,}000-j9{,}000$$

$$=14{,}450-j4{,}020.$$

Expressed in the polar form,

$$Z_t=15{,}000\angle-15.6°.$$

From equation (1–8), the total circuit current is

$$I_t=\frac{E_g}{Z_t}=\frac{200}{15{,}000}=0.01335 \text{ amperes};$$

from equation (1–7), the voltage across the load is

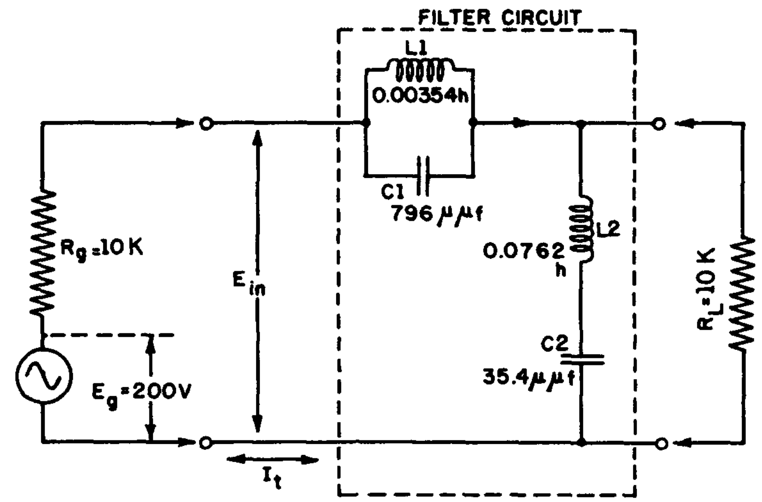
$$E_{3,4}=I_tZ_{3,4}=0.01335\times6{,}670=89.0 \text{ volts};$$

and from equation (1–6), the load current is

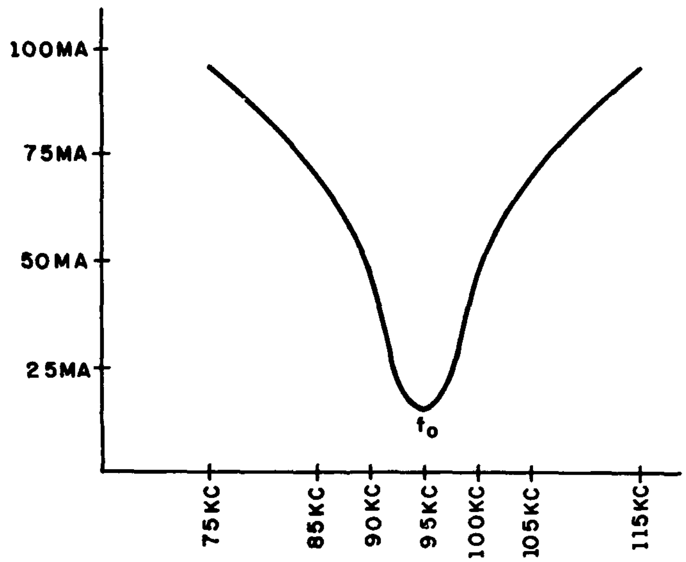
$$I_L=\frac{E_{3,4}}{R_L}=\frac{89}{10{,}000}=0.0089 \text{ amperes}=8.9 \text{ ma}.$$

Band-Elimination Filters

A band-elimination filter (or band-suppression filter) is designed to suppress current of all frequencies within a continuous band, limited by the lower and upper cut-off frequencies, and to pass all frequencies below or above that band. A simple band-suppression filter is shown in figure 1–7, A. This type of filter is just the opposite of the band-pass filter; currents of frequencies within the band are greatly attenuated or weakened. The series- and parallel-tuned circuits are tuned to the center of the band to be eliminated. The parallel-tuned circuit in series with the source offers a high impedance to this band of frequencies, and the series-tuned circuit in shunt with the load offers very low impedance; therefore, the signals within the elimination band are both blocked and diverted from the load. All other currents—that is, currents at all frequencies outside the band—pass through the parallel circuit with very little



A
CIRCUIT



B
CHARACTERISTIC CURVE

Figure 1-7.—Band-elimination filter.

opposition and are unaffected by the series-tuned circuit since it acts as an open circuit at these frequencies.

Assume that a band of frequencies extending from 90 kc to 100 kc is to be suppressed by the filter. For an input and output resistance of 10,000 ohms, the values of inductance and capacitance are as indicated in the figure. (The formulas by which these values are determined may be found in radio engineering handbooks.)

At resonance (95 kc) the parallel circuit, $L1C1$, offers maximum impedance and may be considered as almost an open circuit. At the same frequency the series circuit, $L2C2$, in effect short-circuits the load, so that minimum current will flow through the load at the resonant frequency. As in the case of the band-pass circuit, the inherent series resistance of the two tuned circuits is small and can be neglected.

Below resonance—for example, at 90 kc—the impedances of $L1C1$ and $L2C2$ are such that with an assumed voltage of 200 volts about 42.5 milliamperes flow through the load, R_L. At 75 kc (20 kc below resonance) the current through the load is increased to approximately 96 milliamperes.

The band-suppression characteristic is symmetrical about the resonant frequency and the same relative increase in current with increase in frequency may be assumed. Figure 1–7, B, is an indication of the current-frequency characteristic of the filter shown in figure 1–7, A. The current is the current that flows through the load, R_L.

Wave Traps

Wave traps, sometimes used in the antenna circuits of radio receivers, are forms of band-elimination filters. There are two general types of wave traps—the parallel-tuned filter and the series-tuned filter.

The parallel circuit, in series with the antenna in figure 1–8, A, is tuned to resonance at the frequency of the undesired signal.

The parallel wave trap presents a high impedance to cur-

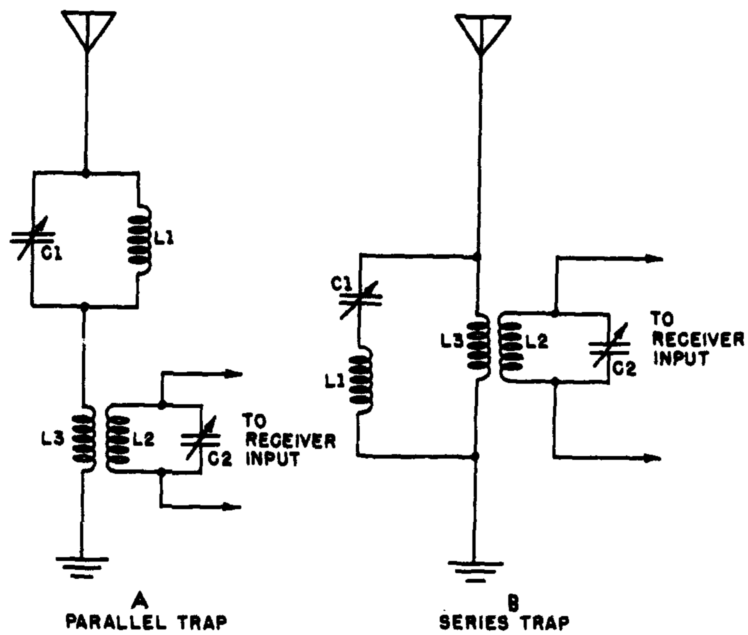


Figure 1-8.—Wave traps.

rents of this unwanted frequency and allows currents of other frequencies to enter the receiver with only slight attenuation.

The series circuit, connected as shown in figure 1-8, B, is tuned to resonance at the frequency of the undesired signal. The impedance of the series circuit, $C1L1$, at resonance is low. Hence, these unwanted frequencies will be bypassed to ground around the receiver input transformer primary, $L3$. The desired frequencies will be essentially unaffected because either L, or C, act as a high impedance when not in resonance.

INDUCTIVELY COUPLED TUNED CIRCUITS

When two separate circuits are so positioned that energy from one circuit is coupled to the other circuit by transformer

action, they are said to be INDUCTIVELY COUPLED. MUTUAL INDUCTANCE is the common property of the two circuits that determines, for a particular rate of change of current in one of the circuits, the amount of emf induced in the other circuit. Mutual inductance is expressed in henries and is designated by the letter M.

Although problems involving inductively coupled circuits may be somewhat complex, they may be simplified if the following assumptions are made: (1) The effect of the presence of the coupled secondary on the primary is the same as if an impedance, $\frac{(\omega M)^2}{Z_s}$, called the COUPLED IMPEDANCE, had been added in series with the primary. In this expression, M is the mutual inductance, and Z_s (a vector) is the series impedance of the secondary, when not coupled to the primary. (2) The secondary voltage, e_2, induced by the primary current, i_p, has a value of $\omega M i_p$ and lags the primary current by 90°. (3) The secondary current, i_s, is that value of current that would flow if the primary were removed and the induced secondary voltage were applied in series with the secondary winding. The coupled impedance is a vector quantity. It has the same phase angle as the secondary impedance, Z_s, but is of opposite sign.

The primary current, secondary voltage, and secondary current may be determined by Ohm's law as applied to a-c circuits. Thus, the primary current, i_p, may be determined as

$$i_p = \frac{E}{Z_p + \frac{(\omega M)^2}{Z_s}},$$

where E is the voltage applied to the primary. The secondary voltage has been stated as $\omega M i_p$. The secondary current, i_s, is determined as

$$i_s = \frac{\omega M i_p \angle \pm 90^\circ}{Z_s}.$$

From the expression for coupled impedance, $\frac{(\omega M)^2}{Z_s}$, some

of the characteristics of a coupled circuit can be determined. For example, if M is large and Z_s is small, the coupled impedance will be large and the primary current may be reduced as a result of the increased impedance when the coupling occurs. The voltage induced in the secondary, and the secondary current, will be correspondingly affected. If, on the other hand M is small and Z_s is large, the effect of the presence of the secondary on the primary is slight and little change occurs in primary current with increased coupling. A special case in which the secondary is tuned is of importance because it is widely used in r-f voltage amplifiers in radio receivers.

Untuned-Primary Tuned-Secondary Circuit

A simplified untuned-primary tuned-secondary circuit is shown in figure 1–9, A.

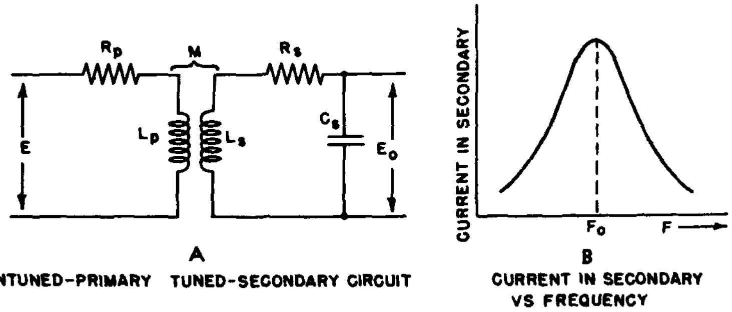


Figure 1–9.—Untuned-primary tuned-secondary circuit and response curve.

Ordinarily, if an electron tube is connected in the primary circuit, the plate resistance, r_p, acts in series with the primary, L_p, and the coupled impedance.

If the coupling is small and the plate resistance is large with respect to the coupled impedance (as in the case in pentode amplifiers), the primary current is substantially independent of the resonant condition of the secondary. The curve of secondary current vs frequency (fig. 1–9, B) then has the same general shape as that of the ordinary series-tuned circuit of figure 1–4. If a triode is used in place of the pentode, the plate resistance is reduced and the coupled im-

pedance becomes an appreciable part of the primary impedance. At resonance the secondary impedance is low and resistive and the coupled impedance causes a dip in primary current. At frequencies away from resonance the secondary impedance increases, and the coupled impedance is less, causing an increase in primary current. This increases the voltage induced in the secondary by transformer action at these off-frequency points and prevents the secondary current from falling off as rapidly as it would if the secondary induced voltage were constant. Thus, the curve is broader and indicates a lower effective Q for the series-tuned circuit than would exist if the plate resistance of the electron tube were removed from the circuit.

The formula of the coupled impedance, which is responsible for the shape of the characteristic curve of an untuned-primary tuned-secondary circuit, is

$$\text{coupled impedance} = \frac{(\omega M)^2}{Z_s} = \frac{(\omega M)^2}{R_s + j\omega L_s - j\dfrac{1}{\omega C_s}}.$$

In the vicinity of resonance, $(\omega M)^2$ varies only *slightly*. The denominator represents the series impedance of the secondary. The coupled impedance formula, $\frac{(\omega M)^2}{Z_s}$, is similar in form to that of the impedance formula of a parallel resonant circuit, $\frac{(\omega L)^2}{Z_s}$. Thus, the coupled impedance due to the tuned secondary varies with frequency according to the same mathematical law as the impedance of a parallel circuit varies with frequency. In the case of coupled impedance, however, the magnitude of the curve depends on the mutual inductive reactance, ωM, instead of the inductive reactance, ωL.

Tuned-Primary Tuned-Secondary Circuit

A simplified tuned-primary tuned-secondary circuit and current vs frequency response curves are shown in figure

1–10, A. As indicated by the response curves (fig. 1–10, B and C), this type of circuit has a band-pass characteristic that depends in part on the coefficient of coupling, k, and in part on the circuit Q's.

Under proper operating conditions essentially uniform amplification of a relatively narrow band of frequencies may be achieved, and amplification of frequencies outside this band may be sharply reduced. These characteristics make this type of coupling highly desirable in intermediate-frequency amplifiers in both radio and television receivers.

Because the slope of the response curve is not perfectly vertical, the circuit cannot completely discriminate against frequencies just outside the desired channel without also attenuating to some extent the frequencies at the upper and lower limits of the pass band. However, double-tuned circuits approach an ideal band-pass characteristic much more closely than do single-tuned circuits, which have rounded response curves.

When the coefficient of coupling (fig. 1–10, B) is low, the response is sharply peaked at the resonant frequency and the pass band is very narrow. As the coupling is increased to the critical value, maximum current flows in the secondary, and the output voltage across the secondary is also at its maximum. At this point (critical coupling),

$$k=\frac{1}{\sqrt{Q_pQ_s}};$$

and if the Q's are equal,

$$k=\frac{1}{Q}.$$

The pass band is still relatively narrow.

If the coupling is further increased until the OPTIMUM value is reached, the gain is still relatively high; but the pass band has been increased and the response is essentially uniform. At this point (optimum coupling),

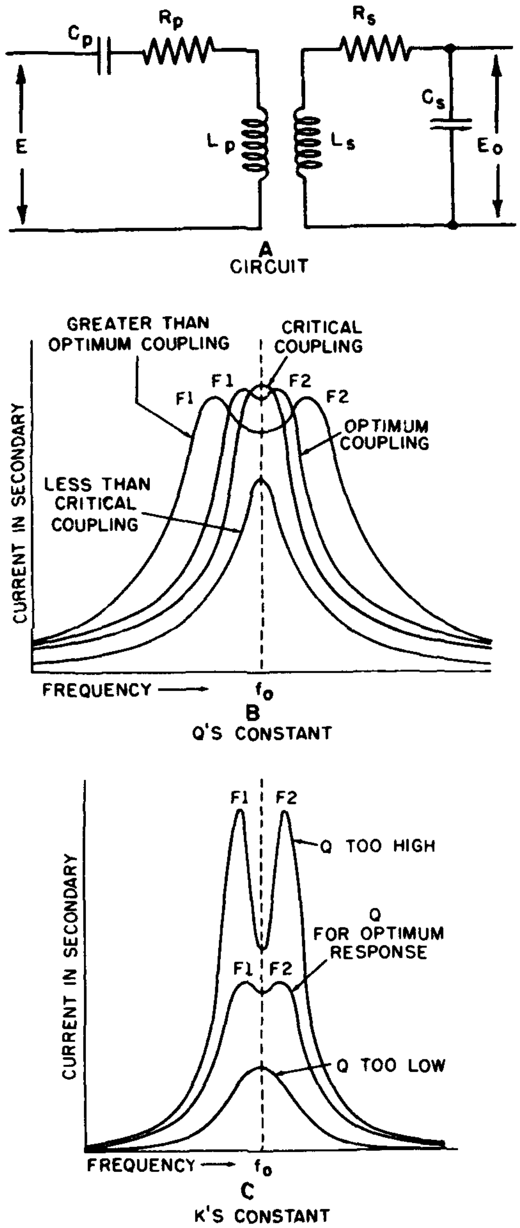


Figure 1-10.—Tuned-primary tuned-secondary circuit and response curves.

$$k = \frac{1.75}{\sqrt{Q_p Q_s}};$$

and if the Q's are equal,

$$k = \frac{1.75}{Q}.$$

As the coupling is again increased, the humps at f_1 and f_2 are well defined, and the gain at resonance is considerably reduced. Although the pass band is now much wider, the gain throughout the band is not sufficiently uniform.

The two humps, f_1 and f_2, in the curve are due to the reactance, $\frac{(\omega M)^2}{Z_s}$, that is coupled into the primary as the coupling is increased. Below resonance this reactance is inductive, and above resonance it is capacitive. For the same frequencies, the coupled reactance has the opposite sign to that of the primary, and the impedance of the primary is therefore reduced. Accordingly, there is an increase in primary current at frequencies slightly off resonance; and this results in increased current in the secondary, and also an increase in voltage at the output.

The frequencies at the two humps, f_1 and f_2, which define the practical lower and upper limits of the pass band, are determined by the following equations:

$$f_1 = \frac{f_o}{\sqrt{1+k}}$$

$$f_2 = \frac{f_o}{\sqrt{1-k}}$$

Figure 1–10, C, shows the effects of varying the Q while maintaining a constant coefficient of coupling. Actually, the desired response curve could be achieved by the proper manipulation of both k and Q because they are interrelated.

From the foregoing equations it is seen that in order to have a wide pass band k must be large, and the circuit Q's must be small. However, the proper relation between k and

Q is essential if both the desired bandwidth and the desired response within the band are to be maintained.

The application of inductively coupled circuits in electron-tube amplifiers, together with amplifier voltage-gain equations, is treated in chapter 5.

QUIZ

1. Why is the reactance of a series-tuned circuit zero at the resonant frequency?
2. What is the most important characteristic of a series-tuned circuit?
3. Under what circumstances will the voltage appearing across either the inductor or the capacitor in a series circuit be much higher than the source voltage?
4. Why is the impedance offered by a parallel-resonant circuit maximum and purely resistive at the resonant frequency?
5. Under what circumstances will an inductor and capacitor be in resonance at the same frequency irrespective of whether they are connected in series or in parallel?
6. What are two of the principal functions of tuned circuits in receivers?
7. What distinguishes a vector quantity from a scalar quantity?
8. How many times must the $+j$ operator be applied as a multiplying factor to a vector initially in the 0° position to rotate it 450°?
9. If a vector, initially at the 0° position (along the $+X$ axis), is multiplied by j^3 and then by $-j^4$ what will be its angular position with respect to the X axis?
10. What magnitude and angle are represented by the expression $j^2 5$?
11. What does the expression $+j^5$ ohms represent?
12. (a) Add: $6+j3$ to $1+j2$
    (b) Subtract: $1+j2$ from $6+j3$ (and reduce to simplest form)
13. (a) Multiply: $3+j^2 4$ by $3-j^2 2$
    (b) Divide: $6+j3$ by $3-j2$
14. Convert from one form to the other:
    (a) $2\angle+30°$
    (b) $6+j8$

15. When may polar vectors be added or subtracted algebraically?
16. (a) Multiply: $3 \angle +30°$ by $6 \angle -60°$
(b) Divide: $4 \angle +40°$ by $2 \angle -20°$
17. Why is the power factor unity in a series-resonant circuit?
18. What determines the average power dissipated in a series-resonant circuit?
19. What two ratios may be used to determine the Q of an inductor?
20. What two ratios may be used to determine the *Q* of a capacitor?
21. How does a decrease in the effective resistance acting in series with a resonant circuit affect the circuit *Q*?
22. How does an increase in the circuit *Q* of a series-resonant circuit affect the voltage gain?
23. How does the circuit *Q* affect the sharpness of the current-vs-frequency curve of a series-resonant circuit?
24. Why does the current in a series-resonant circuit lag the applied voltage above the resonant frequency?
25. What is the principal use of the series-resonant circuit?
26. In a parallel-resonant circuit, why is the line current in phase with the applied voltage?
27. What determines the magnitude of the impedance of a parallel-resonant circuit?
28. In a parallel-resonant circuit why is the phase angle between *R* and *Z* negative above resonance?
29. What are the effects of increasing the load on a parallel resonant circuit?
30. Why is a parallel-resonant circuit called a tank circuit?
31. In the band-pass filter of figure 1–6, what prevents the unwanted frequencies from reaching the load?
32. In the band-elimination filter in figure 1–7 what prevents the unwanted frequencies from reaching the load?
33. What is the function of wave traps?
34. In transformer action what effect does coupled impedance (added in series with the primary) have on the primary current if the source voltage is constant?
35. In an untuned-primary tuned-secondary transformer-coupled circuit, when is the primary current independent of the resonant condition of the secondary?
36. In a tuned-primary tuned-secondary transformer-coupled circuit, what two factors determine the band-pass characteristic?

CHAPTER 2

OPERATING PRINCIPLES OF THE ELECTRON TUBE

The electron tube is considered primarily responsible for the rapid evolution of electronics to its present stage. It is one of the basic components of almost every piece of electronic equipment. Without the discovery and development of the electron tube, elaborate yet compact equipments, such as radio, radar, and sonar would not have been possible. It should, therefore, be apparent to the student that, in order to have a clear concept of electronic theory and the operating principles of electronic equipment, electron tube principles are of utmost importance.

The electron tube is made up of a highly evacuated glass or metal shell which encloses several elements. The elements consist of the cathode, the plate, and sometimes one or more grids. Electron tubes are of many types and designations and perform many functions. They can be made to (1) convert currents and voltages from one waveform to another, (2) amplify weak signals with minimum distortion, and (3) generate frequencies much higher than any conventional a-c generator.

TYPES OF EMISSION

Electrons flow within a conductor when a potential difference is applied across the terminals of the conductor. These electrons break away from the orbits of their parent atoms

and move with a rapid vibratory motion, the velocity of which increases with temperature. At ordinary temperatures the particles do not leave the surface of the conductor because their velocity is not great enough to overcome the attractive forces within the conductor.

To escape from a metallic surface, electrons must do work to overcome the forces of attraction which are always present. This amount of work is called the WORK FUNCTION of the material. Increasing the heat intensity of a metallic emitter increases the kinetic energy of the so-called free electrons in the material.

Thermionic Emission

Thermionic emission is the process by which electrons gain enough energy by means of heat to be released from the surface of the emitter. Thermionic emission is the type of emission most frequently employed in electron tubes.

Photoelectric Emission

An emission of electrons can also be caused by light striking the surface of certain materials. This type of emission is called PHOTOELECTRIC EMISSION. The energy of the light rays striking the substance is imparted to electrons on the surface. If the energy acquired by the electrons is sufficient, the force thus acquired will overcome the attractive forces at the surface and the electrons will escape from the substance. The velocity at which the electrons are emitted is directly proportional to the light frequency of the radiant energy striking the material; therefore, the higher the light frequency (shorter the wavelength) the greater is the velocity of emitted electrons. The number of electrons emitted is directly proportional to the intensity of the light. Materials that are particularly sensitive to light are zinc, potassium, and the other alkali metals. Two of the principal uses for photoelectric emission are photoelectric cells and television camera, or iconoscope, tubes.

Secondary Emission

Emission of electrons from a body caused by the impact of other electrons striking its surface is called SECONDARY EMISSION. If a stream of electrons flowing at a high velocity strikes a material, the force may be great enough to dislodge other electrons on the surface. Secondary emission is not commonly used as a source of electrons. However, it does occur spontaneously in tubes and must be controlled. This problem is discussed later in this chapter.

TYPES OF EMITTERS

Only a few substances can be heated to the high temperatures that are required to produce satisfactory thermionic emission without melting. Tungsten, thoriated-tungsten, and oxide-coated emitters are the only types that are commonly used in electron tubes.

Tungsten Emitters

Tungsten has a great durability as an emitter but requires a large amount of heating power and a high operating temperature for satisfactory emission. Tungsten cathodes are used primarily in high-power electron tubes like those in HIGH-POWER radio transmitting equipment.

Thoriated-Tungsten Emitters

A thoriated-tungsten emitter has a thin layer of thorium on the surface of the tungsten. The layer of thorium is monomolecular—that is, only 1 molecule thick. Thoriated-tungsten cathodes have greater electron emission at a lower operating temperature than a cathode of pure tungsten and are normally used in tubes that are operated at plate voltages of 500 to 5,000 volts. Tubes such as the 860 and 861 use this type of emitter. These tubes and others like them are used extensively in low-power radio transmitters.

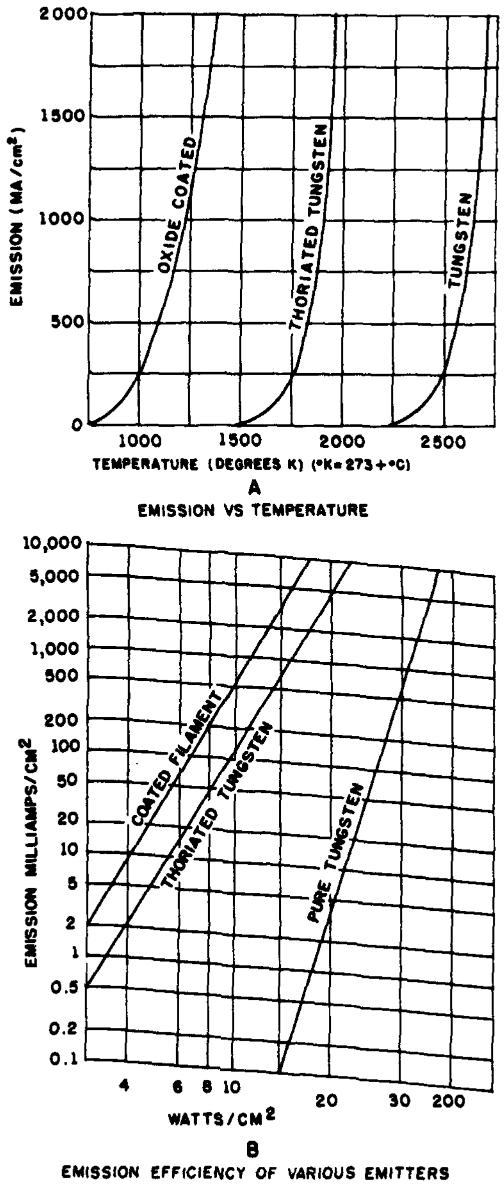


Figure 2-1.—Emission vs temperature curves for three types of emitters.

Oxide-Coated Emitters

Oxide-coated emitters consist of metal, such as nickel, coated with a mixture of barium and strontium oxides, over which is formed a monomolecular layer of metallic barium and strontium. This is the most efficient type of emitter. It operates at a lower temperature than tungsten or thoriated-tungsten and therefore requires less power, resulting in a longer life at a higher emission efficiency. It is used in almost all types of receiving tubes.

The graphs in figure 2–1, A, show electron emission as a function of cathode temperature for the three types of emitter materials discussed in this chapter. The temperature at which emission becomes appreciable is called the NORMAL OPERATING TEMPERATURE. The emitter comprises the cathode of the electron tube. The emission efficiency of the three types of emitter materials is shown in figure 2–1, B. (See page 51.)

HEATING THE EMITTER

The electron-emitting cathodes of electron tubes are heated in two ways—(1) directly, and (2) indirectly. A directly heated emitter receives its heat by the passage of a current through the filament itself which serves as the cathode. An indirectly heated cathode comprises a metal sleeve that surrounds the filament but is electrically insulated

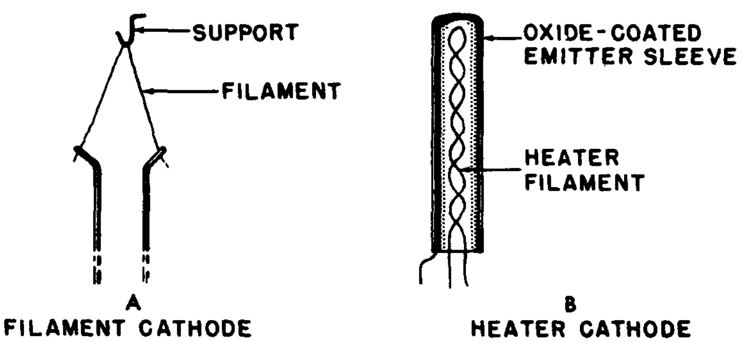


Figure 2–2.—Methods of heating the cathodes in electron tubes.

from it. The sleeve serves as the cathode emitter and receives its heat mostly by radiation. Both types are shown in figure 2–2.

Directly heated cathodes are generally employed in portable equipment that is supplied from batteries. The d-c filaments of these tubes are so constructed that the drain on the filament battery will be low. Indirectly heated cathodes would require too much power for heating purposes. Because the filament current is steady, the heating is uniform; and the filament cross section is relatively small compared with a-c filaments. Directly heated a-c filaments require relatively large cross sections to reduce the temperature variations that occur at twice the power frequency. When a-c power is available, it is common practice in receiving equipment to employ indirectly heated cathodes. The cathode in this type of tube is isolated from the a-c heater supply, and therefore hum occurring at the power frequency (or at twice the power frequency) is largely eliminated.

PHYSICAL CHARACTERISTICS OF ELECTRON-TUBE MATERIALS

The outer walls of an electron tube are constructed either of thin glass or metal. The larger the tube, the thicker the glass must be because of the greater weight of the atmosphere to be sustained on the walls of the tube. The physical size of a tube is determined by (1) performance function, and (2) average designed equipment space in which the electron tube is to be used.

Evacuation of air from a tube is required for two reasons—(1) to prevent destruction of the cathode and heating element by oxidation or burning, and (2) to allow the flow of current from cathode to plate without colliding with gas particles. The lightest gas particle is approximately 1,800 times as heavy as an electron; thus a gas molecule would divert an electron upon impact and make the current flow erratic.

High vacuum is produced by burning a small amount of

magnesium or barium, known as a "getter," inside the tube after it has been sealed off from the outside air and after most of the air has been removed with high-vacuum pumps. The getter is ignited by means of a high-frequency coil which is placed around the electron tube. The high-frequency field induces eddy currents in the metal within the tube. These eddy currents heat up a metal cap which contains a small charge of gun powder. This heat fires the magnesium getter and combines with gas left in the tube to form a silvery deposit on the inner walls. This deposit of magnesium carbonate occupies much less space than did the gas, hence the degree of vacuum is increased.

The external leads from the tube are electrically welded to the tube elements and brought out at the bottom through a special glass-metal fusion to make the envelope airtight. In metal tubes a glass button is used at the base to afford electrical insulation. The materials selected for the external leads have nearly the same coefficient of expansion as that of glass. Thus during heating and cooling periods, the glass expands and contracts the same amount as the metal and the vacuum seals are maintained.

Metal tubes are an outgrowth of the competitive field in tube manufacturing. They are designed to act as a shielded unit (the same as a glass tube with an external shield placed over it). However, during the period of World War II when metals were scarce, the GT series tubes were manufactured and found to be very satisfactory. The shield on a tube acts primarily to prevent the introduction of stray fields within the envelope where induced voltages might be amplified many times, thus causing distortion in the output stages. There are a few circuits in electronic equipments where metal and glass tubes cannot be interchanged. Before making an interchange, the technician should always investigate.

The spacing of the electrodes in a tube is dependent on many factors but the two most important are (1) frequency utilization and (2) interelectrode voltages.

The anode (plate) is made of materials that will not emit

electrons by thermionic means at normal tube operating temperatures. Metals used as plates include iron, nickel, carbon, and tantalum. The plate is mounted externally with respect to the cathode. It is electrically insulated from the cathode and usually surrounds it in order to receive all of the cathode field of emission. The plate usually has a dark surface to radiate the heat caused by the plate current.

Electron tubes are identified by a number or a combination of numbers and letters. So many different types of tubes have been introduced that it has become impossible to adhere rigidly to the system as it was originally set up. However some of the ideas contained in the original system have been followed for many years. In the old system, the type number is divided into four parts. First, a number consisting of one or more digits designates the filament or heater voltage. Second, one or more letters designate the type or function of the tube. Third, a number designates the number of useful elements in the tube. Fourth, one or more letters designate the size or construction. For example, the type 6SK7 electron tube is a variable-mu pentode having a filament voltage of approximately 6 volts and featuring single-ended construction (no grid cap) and inter-lead shields. Because thousands of different types of receiving tubes are manufactured there are probably more exceptions to this system of designation than there are tubes that follow it completely. It is therefore desirable to refer to a tube manual when a tube characteristic is in question.

DIODES

The simple 2-element tube contains a cathode and plate. The plate collects electrons emitted from the cathode and provides a connection to the external circuit. Figure 2–3, A, shows a side cut-away view of a filament cathode type of diode and figure 2–3, B, shows a similar view of a separate heater type of diode.

In a circuit the tube acts in the manner of a valve (actually called a valve by the British instead of tube). The

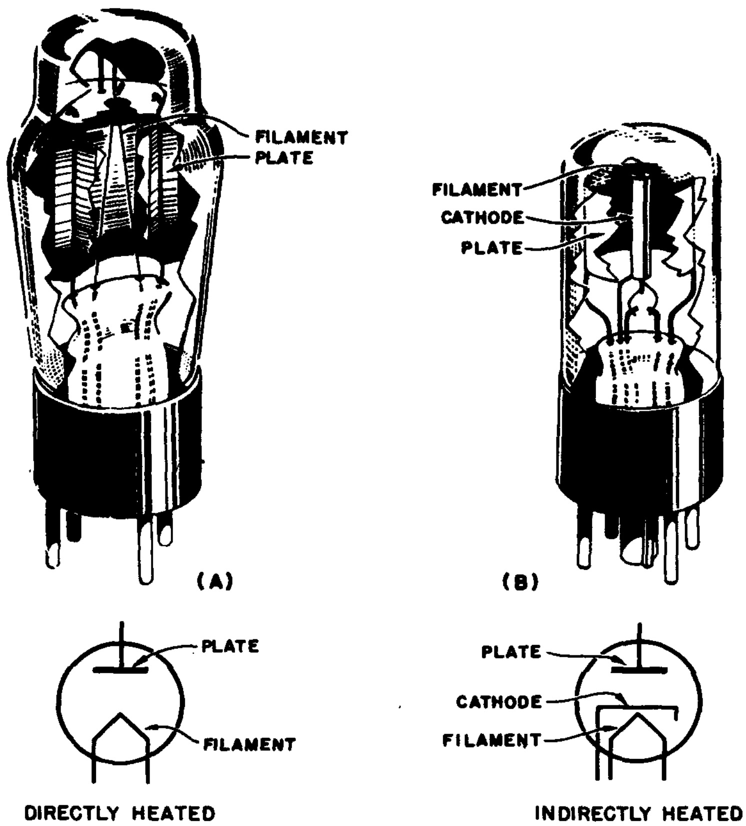


Figure 2-3.—Cut-away of 2-element tubes.

opening and closing of the valve occurs as the positive and negative voltages are applied to the tube. This effect may be observed by connecting the plate circuit in series with a battery and milliammeter and returning it to the cathode (fig. 2-4). The cathode is brought up to normal operating temperature by applying rated voltage across the heater terminals. If the battery is connected so that the plate is positive with respect to the cathode (fig. 2-4, A), the meter will indicate a current flow. This phenomenon, first ob-

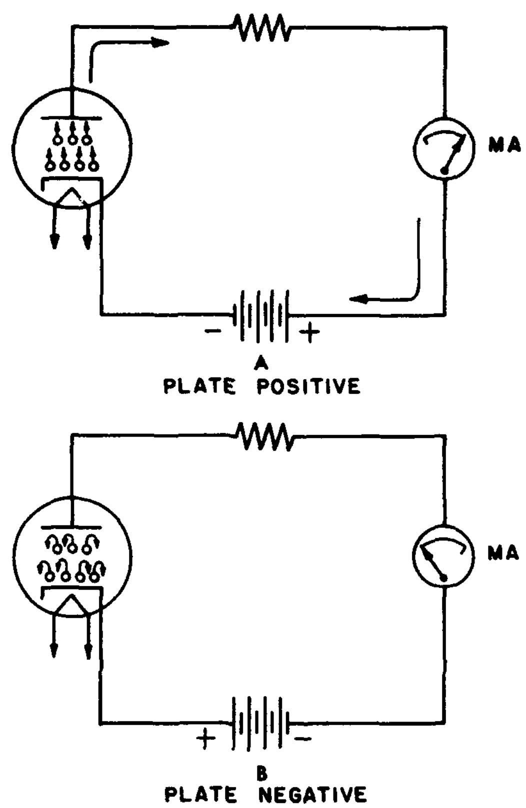


Figure 2-4.—Action of diode.

served by Edison, is known as the EDISON EFFECT. If the battery is connected so that the plate is negative with respect to the cathode, the meter will indicate no plate current flow.

The circuit current constitutes the simultaneous movement of electrons around the circuit with those that are moving within the tube from the cathode to the plate. Thus electrons are returned to the emitter as fast as they leave it and are removed from the plate as fast as they strike it so that neither cathode nor plate acquires a blocking charge. The diode functions in this way when the battery polarity makes the plate positive with respect to the cathode.

The total number of electrons emitted by the cathode at a given operating temperature is always the same regardless of the plate voltage. The electrons in the space around the cathode constitute a negative space charge that constantly tends to repel the electrons back to the cathode as fast as they are being emitted.

At low plate voltage only those electrons nearest the plate are attracted to it and the plate current is low. As the plate voltage is increased (the cathode temperature remaining constant), greater numbers of electrons are attracted to the plate and correspondingly fewer of those being emitted are repelled back into the cathode.

Eventually a plate voltage (SATURATION VOLTAGE) is reached at which all the electrons being emitted are in transit to the plate and none are repelled back to the cathode. The corresponding value of current is called the SATURATION CURRENT. Any further increase in plate voltage can cause no further increase in the plate current flowing through the tube.

The relation between the plate current in a diode and the plate potential for different cathode temperatures for oxide-coated, tungsten, and thoriated-tungsten cathodes is shown in figure 2–5. At high plate voltages the flow of plate current is practically independent of plate voltage but is a function of the cathode temperature. However, at lower values of plate voltage the plate current is controlled by the voltage between the plate and cathode and is substantially independent of the cathode temperature.

In other words, with a fixed plate voltage, electron emission and plate current will increase with cathode temperature until at some value of temperature the plate current is limited by the space charge. Thus more electrons are being emitted by the cathode than are being attracted by the plate. Continued increase of cathode temperature fails to produce any further increase in plate current. The temperature at

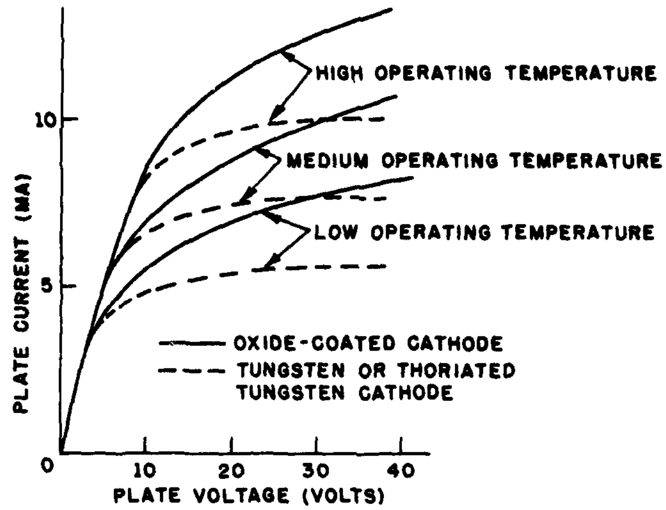


Figure 2-5.—Diode plate-current plate-voltage characteristic curves for various operating temperatures.

which the plate current stops increasing is called the SATURATION TEMPERATURE.

The dotted portion of the characteristic curves is representative of tungsten and thoriated-tungsten emitters, and the solid curves are typical of oxide-coated emitters. It is unlikely that the plate current in a tube employing an oxide-coated emitter will ever be entirely independent of the plate voltage. Before the plate voltage could be increased sufficiently to produce EMISSION SATURATION it is probable that the cathode would be damaged seriously.

Types

Diodes that have been discussed thus far are of the high-vacuum type. There are other types of diodes that contain gas at a relatively low pressure. For example, hot-cathode mercury-vapor rectifier tubes are used extensively to provide plate power for large transmitters. In other applications cold-cathode diodes containing gas at low pressure are

used as voltage regulators, relaxation oscillators, and transmit-receive switching devices.

Uses

Since current can flow in only one direction through a diode its basic use is as a RECTIFIER. If the battery in figure 2–4 is replaced with an alternating voltage source, current will flow through the load resistor in the plate lead only on alternate half cycles—when the plate is positive with respect to the cathode. This unidirectional characteristic of the diode is also used in principle when the tube is employed as a DETECTOR.

TRIODES

Construction

The triode, or 3-element electron tube, is similar in construction to the diode, except that a grid of fine wire is added between the cathode and the plate. The addition of the grid gives to the tube its most useful function—the ability to amplify. It is common practice to make the grid in the form of a spiral helix of circular or elliptical cross section with the cathode at the center. Other arrangements, however, may be used provided the essential requirement of being able to control the flow of plate current is met. The space between the meshes is sufficiently large not to block the flow of electrons from cathode to plate. On the other hand, the grid mesh is sufficiently small and close enough to the cathode to control effectively the flow of plate current when the proper voltage is applied between the grid and cathode. The grid is called the CONTROL GRID (G_1) to distinguish it from other grids that are used in multi-element tubes.

The construction features of a typical triode are shown in figure 2–6. Electrical connections to the grid and plate are made through the base pins and support wires. The cathode sleeve is insulated from the filament and is connected by means of a short lead to one of the base pins. The grid is seen to be much closer to the cathode than to the plate.

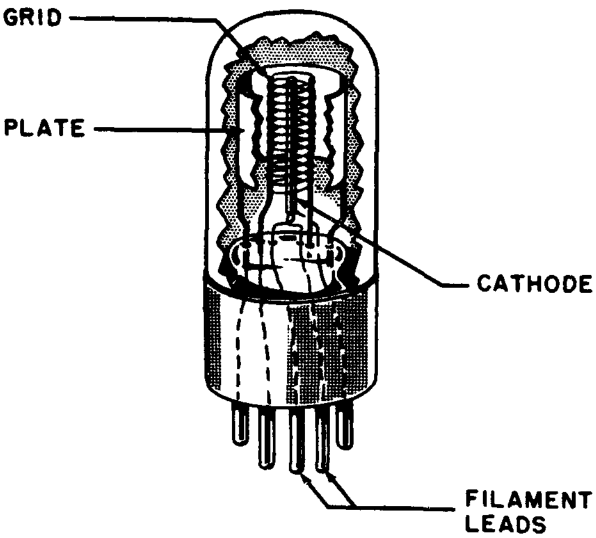


Figure 2–6.—Typical trio.

Operation

Plate current in a given diode depends on the plate voltage and the cathode temperature. Plate current in a triode depends not only on these factors, but also on the grid-to-cathode voltage. A small change in grid voltage causes a relatively large change in plate current. The effective grid control of plate current is caused by its close proximity to the cathode and its placement in a region of the heaviest negative space charge. A small change in grid voltage will produce the same variation in plate current that is produced by a much larger variation in plate voltage. If the grid-to-cathode voltage is increased sufficiently and the grid is negative with respect to the cathode, plate current will stop flowing. The smallest voltage between grid and cathode, with the grid end negative, that will cut off the flow of plate current is called the CUTOFF BIAS.

Figure 2–7 indicates the effect on plate current of making the control-grid voltage progressively less negative with

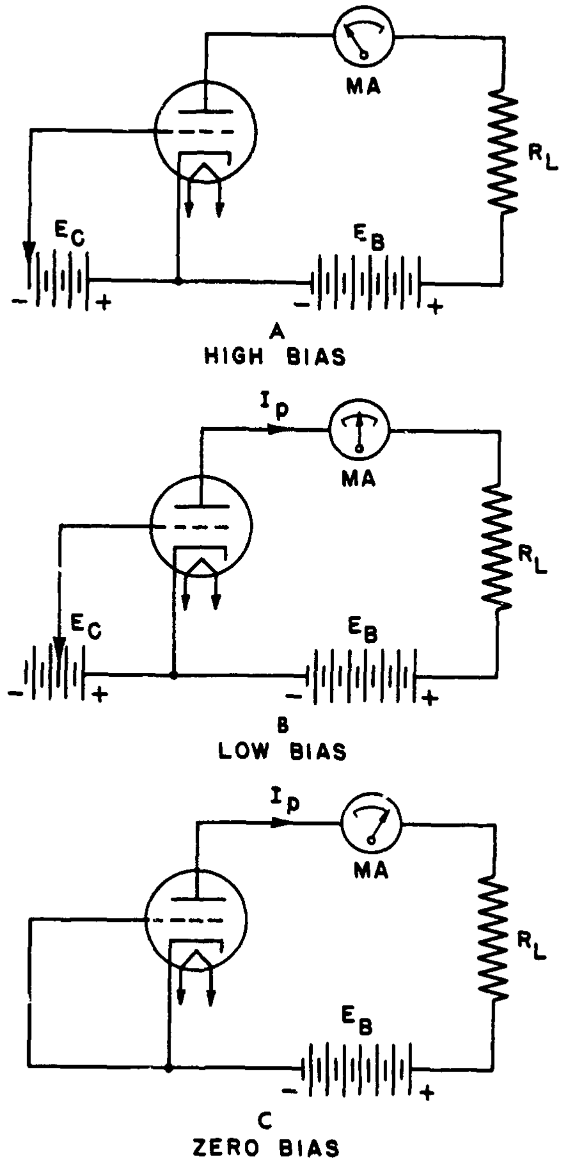


Figure 2-7.—Effect of control-grid voltage on plate current.

respect to the cathode. When the negative bias is high (cutoff or higher), as in figure 2–7, A, no plate current flows because the negative charge on the grid is sufficient to repel the electrons back toward the cathode. As the bias is reduced (fig. 2–7, B) more electrons pass through the grid spaces on their way from the cathode to the plate. With zero bias (fig. 2–7, C) the grid has little or no control on the electron flow to the plate and the triode operates much the same as a diode. As long as the grid is negative with respect to the cathode no grid current flows and no power is consumed in the grid circuit.

If the grid is made positive with respect to the cathode the electrons in the space charge are accelerated toward the plate. Some of them, however, will be attracted to the grid, and grid current will flow, the amount depending on the magnitude of the positive charge on the grid. Power will then be dissipated in the grid circuit. When this power dissipation is undesirable, the grid bias is increased to the point where the peak positive a-c signal voltage will not cause the grid to be positive with respect to the cathode and no grid current will flow.

Amplification

The grid may be considered as an electronic control valve that regulates the flow of electrons through the tube and through the load in the plate circuit. Thus an a-c signal of sine waveform appearing in series with the grid bias causes the plate current to vary in the same manner. The variations in plate current through the plate load are accompanied by corresponding variations in plate voltage. These plate voltage and current variations constitute the output signal of the stage. A relatively small variation in grid input signal is accompanied by a relatively large variation in output signal. Thus the grid signal is said to be amplified in the plate circuit.

Tube Characteristics

The characteristics of electron tubes with cathode, grid, and plate elements involve the relation between grid voltage,

plate current, and plate voltage. Linear and nonlinear characteristic curves can be of either the STATIC or the DYNAMIC type. Both static and dynamic characteristic curves exist for each electron tube. They differ in shape as well as in the actual values they represent. A simple explanation of the difference between these two types of curves is that in static characteristics the values are obtained with different d-c potentials applied between grid, cathode, and plate, and the results are not typical of actual circuit operation. The dynamic characteristics are the values obtained with both a-c and d-c components present as in actual operation. The static characteristics provide an understanding of how the tube itself operates and are discussed in this chapter.

The first characteristic, which is a measure of the voltage amplification of which a tube is capable, is known as the AMPLIFICATION FACTOR, designated μ (pronounced mu). It is the ratio of the increase in plate voltage to the increase in grid voltage required to produce the same change in plate current. An e_p-e_g characteristic for a triode is shown in figure 2-8. A value of plate voltage, e_p, is selected and the grid voltage, e_g, is adjusted to operate the tube at point A on the 20-ma plate-current curve which is arbitrarily selected in this figure.

The value of e_p is raised a specific amount, Δe_p, and e_g is made more negative by an amount, Δe_g, that will hold the plate current constant at 20 milliamperes. The tube now operates at point B. The amplification factor is determined by the ratio of Δe_p to Δe_g, and is expressed as

$$\mu = -\frac{\Delta e_p}{\Delta e_g}. \qquad (i_p \text{ constant})$$

T - minus sign indicates that the changes in plate and grid voltages are in opposite directions. Triodes have practical amplification factors of from 3 to 100.

A second important characteristic is the VARIATIONAL, or a-c plate resistance, designated r_p. It is the ratio, for a constant grid voltage, of a small plate voltage change, Δe_p, to the

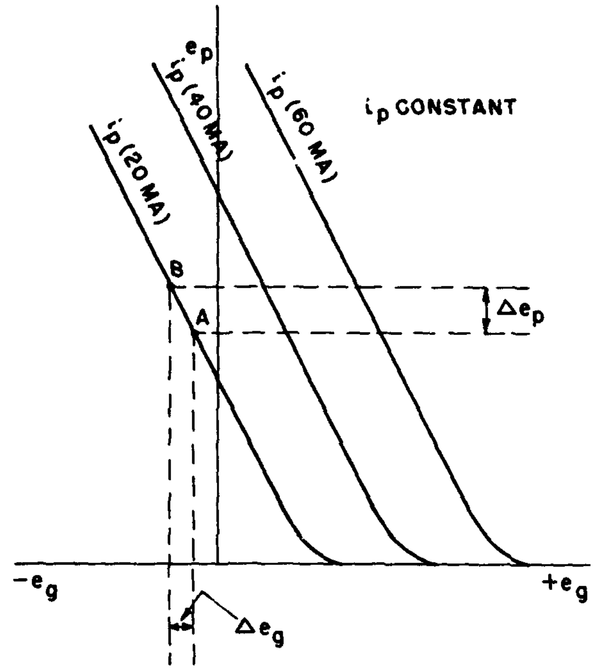


Figure 2-8.—Triode e_p-e_g curves.

resulting small plate current change, Δi_p. It is expressed in ohms when Δe_p is in volts and Δi_p is in amperes. Three i_p-e_p characteristic curves for a triode are shown in figure 2-9. The middle curve is arbitrarily chosen and the grid bias of -2 volts held constant as the plate voltage is adjusted to operate the tube at point A. The plate voltage is increased an amount Δe_p so that the tube now operates at point B. The ratio of this small increase in plate voltage, Δe_p, to the small increase in plate current, Δi_p, which it produces is a measure of the variational, or a-c, plate resistance. Thus,

$$r_p=\frac{\Delta e_p}{\Delta i_p}. \qquad (e_g \text{ constant})$$

A third characteristic used in describing the properties of electron tubes is the grid-plate *transconductance*, designated g_m. It is defined as the ratio, with plate voltage held constant, of a small change in plate current to the small change in grid voltage that causes the change in plate current. It is usually expressed in micromhos. The MHO is the unit of conductivity and is the reciprocal of the ohm, or $\frac{I}{E}$. The word "mho" is "ohm" spelled backward.

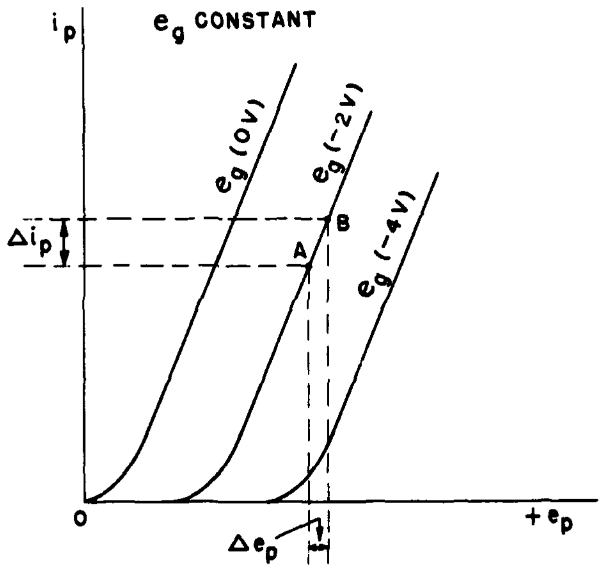


Figure 2-9.—Triode i_p-e_p curve.

Figure 2–10 shows the i_p-e_g characteristics for a triode. In the middle curve of this figure the plate voltage is held constant at 200 volts and the grid voltage is adjusted so that the tube is operated at point A. The grid voltage is reduced an amount, Δe_g, and the tube then operates at point B. The ratio of the small change in plate current, Δi_p, to the small change in grid voltage, Δe_g, indicates the transconductance—that is,

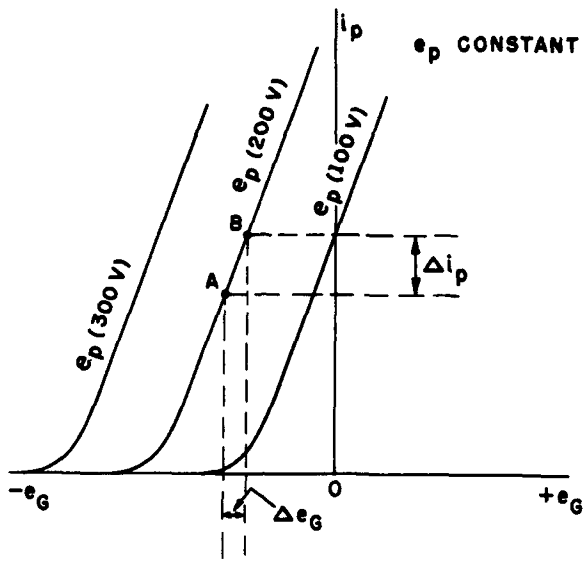


Figure 2–10.—Triode i_p-e_g curve.

$$g_m = \frac{\Delta i_p}{\Delta e_g}. \qquad (e_p \text{ constant})$$

If i_p is expressed in amperes and e_g in volts, g_m must be multiplied by 1,000,000 to express the result in micromhos.

These tube characteristics are interrelated and depend primarily upon the tube structure. This relation is defined by the expression

$$\mu = g_m r_p,$$

where g_m is in mhos and r_p is in ohms.

Distortion

Figure 2–11 illustrates the effects on the output current curve of shifting the bias from a value that allows the tube to operate on the straight portion of the i_p-e_g characteristic curve to a value that forces it to operate on the nonlinear portion of the curve. When the operatin is at point *A*

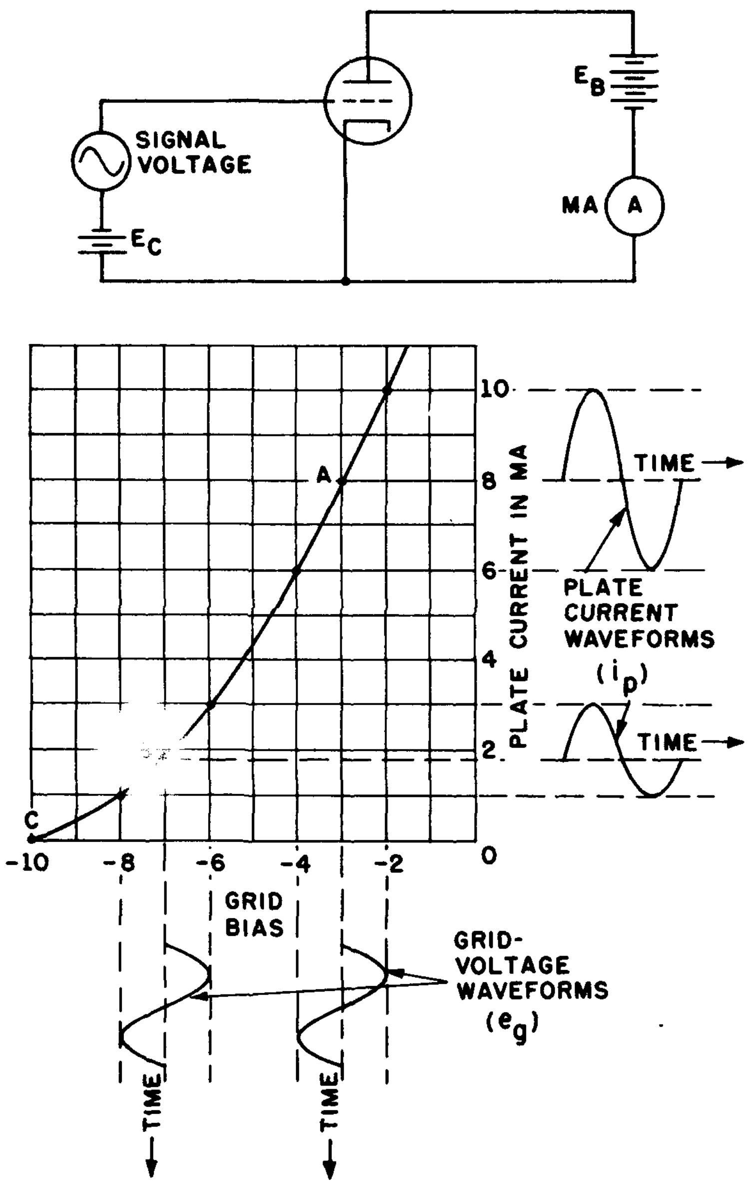


Figure 2-11.—Nonlinear operation due to excessive bias.

(−3 volts), the grid-voltage variation is within the limits of the straight-line portion of the characteristic curve and the plate current faithfully reproduces the grid-voltage waveform. However, if the fixed bias is increased to point *B* (−7 volts), the amplitude of the output waveform is considerably distorted. The extent of this distortion depends upon the actual biasing point of the tube and the extent of the grid-voltage swing.

The point on the zero axis intersected by the characteristic curve (point C in fig. 2–11) is commonly known as the CUTOFF POINT. An amplifier biased to cutoff functions much as a diode rectifier because only alternate half cycles are reproduced in the output circuit. When an amplifier is biased well beyond cutoff and is driven with an excessively large input grid voltage, only that part of the grid-voltage waveform extending into the operating region of the characteristic curve is reproduced in the output. The input signal is thus distorted in the output because only a small portion is amplified. Other forms of distortion are treated in the chapters on electron-tube amplifiers.

Interelectrode Capacitance

Capacitance exists between any two metal surfaces separated by a dielectric. The amount of capacitance depends upon the area of the metal surfaces, the distance between them, and the type of dielectric. The electrodes of an electron tube have a similar characteristic, known as INTERELECTRODE CAPACITANCE, which is illustrated schematically in figure 2–12. The capacitances that exist in a triode are the grid-to-cathode capacitance, the grid-to-plate capacitance, and the plate-to-cathode capacitance.

The shunting effect of the interelectrode capacitance of a tube is increased when the electrodes are connected to a circuit having grid, plate, and cathode leads of appreciable length. The capacitance is increased because of the increase in area afforded by the conducting surfaces comprising the circuit wiring, tube bases, sockets, and so forth.

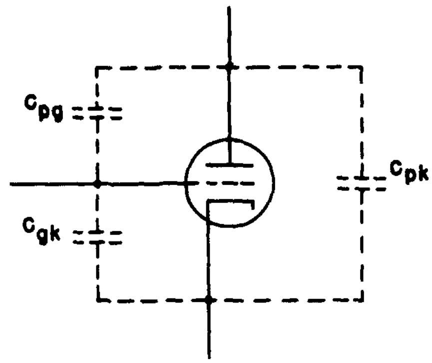


Figure 2-12.—Schematic representation of interelectrode capacitance.

At low and medium frequencies the interelectrode capacitances, as well as the distributed capacitances due to circuit wiring and so forth, have only a slight shunting effect because the reactance at these frequencies is high compared with that of other circuit components.

At high frequencies the interelectrode and distributed capacitances cause appreciable shunting effect because of the reduced reactance offered at these frequencies. Also the grid-to-plate capacitance can feed back some of the plate-signal voltage in the proper phase with respect to the grid-signal voltage to cause undesired oscillations. The effect of this interelectrode capacitive feedback can be neutralized by introduing by means of a capacitor, a voltage of equal magnitude and opposite polarity from the plate to the grid circuit. Such an external capacitor is called a NEUTRALIZING CAPACITOR. It is usually variable to permit adjustment for precise cancellation of the objectionable internal feedback voltage.

At ultrahigh frequencies (u-h-f) interelectrode capacitance becomes very objectionable and prevents the use of ordinary electron tubes. Special u-h-f tubes are used at such operating frequencies. These are characterized by tube elements having very small physical dimensions and spaced electrodes that often do not terminate in conventional tube bases.

MULTIELEMENT TUBES

Many desirable characteristics may be attained in electron tubes by the use of more than one grid. Some common types include TETRODES which contain 4 electrodes, and PENTODES, which contain 5 electrodes. Others containing as many as 8 electrodes are available for certain applications.

Tetrodes

The relatively large values of interelectrode capacitances of the triode, particularly the plate-to-grid capacitance, impose a serious limitation on the tube as an amplifier at high frequencies. Also the relatively low amplification factor makes necessary many stages which accentuate the feedback between plate and grid. To reduce the plate-to-grid capacitance, a second grid called a SCREEN GRID (G_2) is inserted between the grid and plate of the tube, as shown in figure 2–13.

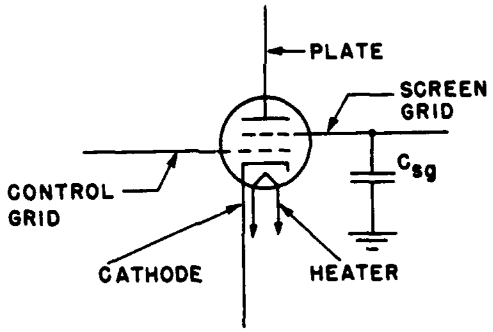


Figure 2–13.—Schematic diagram of a tetrode.

Because the screen grid is shunted by a screen bypass capacitor, C_{sg}, having a low reactance at the signal frequency it acts as a shield or screen between the plate and control grid. It effectively reduces the interelectrode capacitance coupling between the plate and control grid circuits. The screen is supplied with a potential somewhat less positive than the plate. The positive voltage on the screen grid accelerates the electrons moving from the cathode. Some

of these electrons strike the screen and produce a screen current which generally serves no useful purpose. The larger portion, however, passes through the open-mesh screen grid to the plate.

Because of the presence of the screen grid, a variation in the plate voltage has little effect on the flow of plate current. The control grid, on the other hand, retains its control as in the triode. The tetrode has high-plate resistance and an amplification factor ranging up to 800. The high amplification factor is brought about by the close proximity of the control grid to the cathode and the electrical isolation of the plate from the control grid. The transconductance of tetrodes is also relatively high compared with that of triodes.

A typical family of i_p-e_p characteristic curves of a tetrode is shown in figure 2–14, A.

The negative slope of the plate characteristic at plate voltages lower than the screen voltage (90 v) is the result of SECONDARY EMISSION *from the plate*. This condition results from the fact that with the screen voltage fixed, the velocity with which the electrons strike the plate increases with plate voltage. When the electrons strike the plate with sufficient force, other loosely held electrons are knocked out of the plate material into the space between the plate and the screen. Because the screen is at a higher positive potential than the plate, these secondary electrons are attracted to the screen. The flow of these electrons to the screen is in the opposite direction to the normal flow from cathode to plate and the plate current is decreased. This reduction in plate current continues until the potential of the plate approaches the screen-grid potential. Further increase in plate voltage causes the secondary electrons to be pulled back to the plate and the plate current again increases.

The action in the region where plate current decreases as plate voltage increases is called NEGATIVE RESISTANCE. This action is opposite to that encountered in a normal resistor.

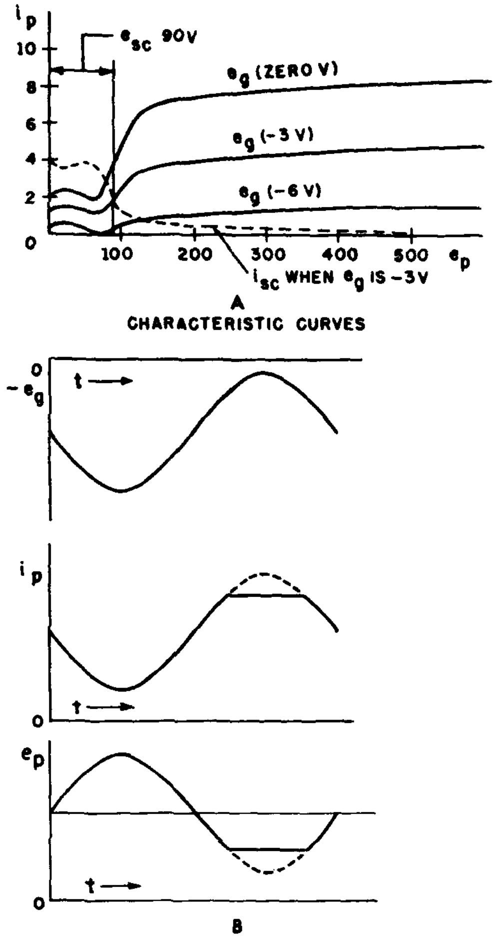


Figure 2-14.—Typical i_p-e_p tetrode characteristic curves.

When the tetrode is used as an amplifier, plate voltage should not fall below the screen voltage. If plate voltage falls below that of the screen, plate current will fail to follow the grid-signal waveform and the output-signal plate-voltage variation is clipped as shown in figure 2-14, B. This distortion may be eliminated by reducing the amplitude of the

grid signal or increasing the B-supply voltage. However, the relatively large screen current and the effects of secondary emission from the plate limit the usefulness of the tetrode as an r-f voltage amplifier.

Pentodes

The effects of secondary emission in the tetrode may be eliminated by the addition of a third grid. The pentode (5-element tube) includes a suppressor grid inserted between the screen grid and the plate for the purpose of suppressing secondary emission from the plate. The 5 elements are: cathode, control grid (G_1), screen grid (G_2), suppressor grid (G_3), and plate. Figure 2–15 is a schematic diagram of a pentode.

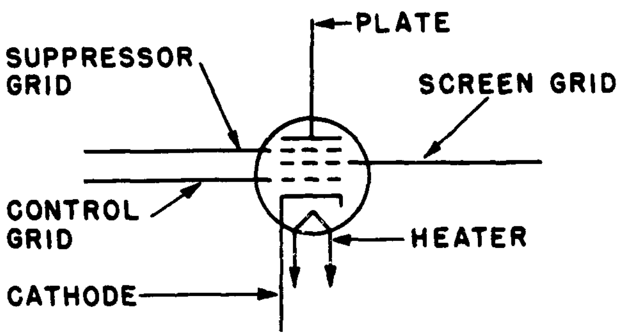


Figure 2–15.—Schematic diagram of a pentode.

In the pentode, the suppressor grid (usually internally connected to the cathode) serves to repel or suppress secondary electrons from the plate. It also serves to slow down the primary electrons from the cathode as they approach the suppressor. These actions do not interfere with the flow of electrons from cathode to plate but serve to prevent any interchange of secondary electrons between screen and plate. The suppressor thus eliminates the negative resistance effect which appears in the tetrode in the region where plate voltage falls below that of the screen.

Thus plate current rises smoothly from zero up to its saturation point as plate voltage is increased uniformly with grid voltage held constant. Typical pentode i_p-e_p characteristic curves are shown in figure 2–16.

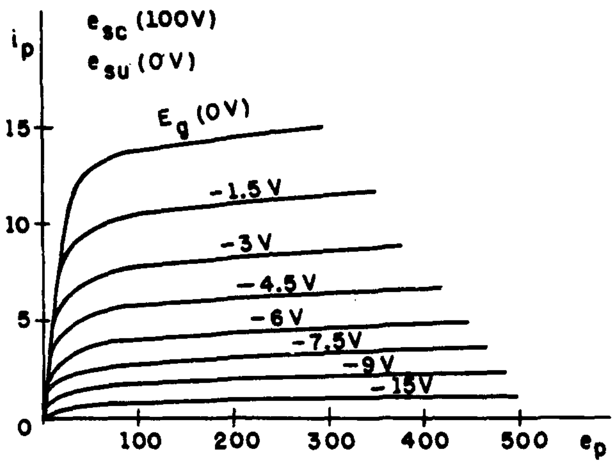


Figure 2–16.—Pentode i_p-e_p characteristic curves.

Pentodes produce an increased voltage-output signal for a given input-grid voltage compared with triodes. The amplification factor of pentodes ranges up to 1,500. The plate resistance and transconductance are both high. In the r-f pentode the chief purpose of the screen grid is to eliminate the effects of interelectrode capacitance coupling between control grid and plate circuits. In the power pentode, at audio frequencies, the screen permits the output signal plate voltage variation to be relatively large without the degenerative action occurring in the triode. Plate current is substantially independent of plate voltage in the power pentode since the screen voltage is the principal factor influencing plate current. With the addition of the suppressor the allowable output voltage variation is larger than that of the tetrode and the distortion effects shown in the

tetrode of figure 2–14, B, are eliminated. Thus an audio-frequency power pentode has an allowable output voltage variation in which the plate voltage can fall a large amount below that of the screen voltage on the positive half cycle of input signal without clipping the plate signal current. Thus the ratio of output power to grid driving voltage is relatively large.

Beam-Power Tubes

A beam-power tube is so named because it is constructed so that the electrons flow in concentrated beams from the cathode through the grids to the plate. The only difference in construction between the beam-power tube and a normal tetrode and pentode is that in the beam-power tube the spaces between the turns of the grids are lined up, and two beam-forming plates are added.

The internal structure of a beam-power tube is shown in figure 2–17. Because the spaces between the grids are lined

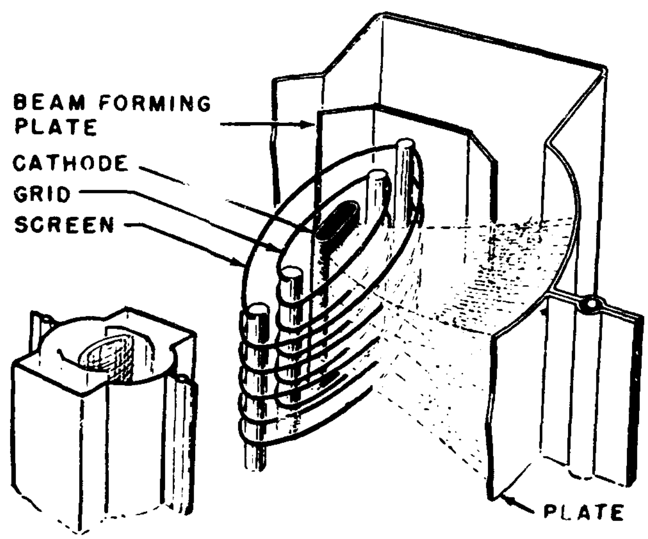


Figure 2–17.—Internal structure of a beam-power tube.

up, fewer electrons strike the screen grid; therefore, screen grid current is lower and plate current higher than in other pentodes. Furthermore, when no actual suppressor is used, the beam-forming plates at cathode potential produce the desired beam effect. Secondary emission from the plate is then reduced because of the space charge between the screen grid and plate.

The space charge results from the slowing down of the electrons as they pass from the high-potential screen to the lower potential plate. The space charge thus formed in front of the plate is sufficient to repel back to the plate secondary electrons emitted as a result of the impact of primary electrons. This action also increases plate current and reduces the screen current.

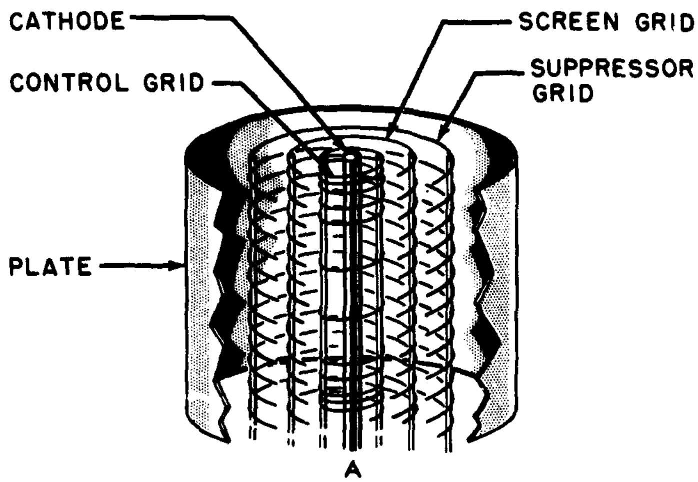
A beam-power tube that is operated at the same plate and screen voltages as a normal tetrode provides more power output for a given signal voltage with no increase in internal tube capacitances.

Variable-Mu Tubes

The amplification of a tube may be controlled by varying the bias voltage applied to the grid, but normally the range of this control is limited by the cutoff bias and the permissible distortion. In receivers employing automatic volume control (a-v-c) in the r-f amplifier section, the amplification is varied over a wide range so that strong or weak signals may be accommodated. To permit this increased range of volume control, the VARIABLE-MU tube was developed. This tube is also known as the SUPER CONTROL or REMOTE CUTOFF type.

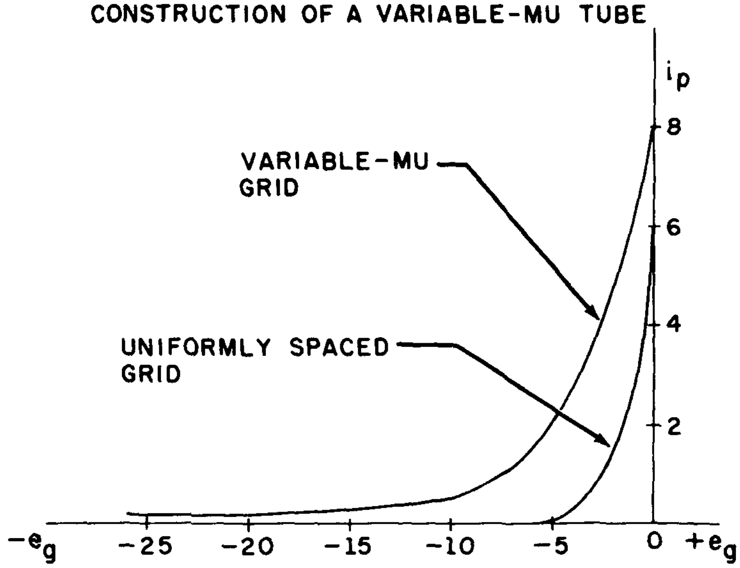
The only difference in construction between variable-mu tubes and normal, or sharp cutoff tubes is in the spacing between the turns of the control grid. In sharp cutoff tubes the turns of the grid wires are equally spaced, while in remote cutoff types the grid turns are closely spaced at the ends and widely spaced in the center. The construction of variable-mu tubes is shown in figure 2–18, A.

With a small bias voltage, electrons flow through all the



A

CONSTRUCTION OF A VARIABLE-MU TUBE



B

$i_p - e_g$ CURVE FOR NORMAL AND VARIABLE-MU TUBES

Figure 2-18.—Construction of variable-mu tubes and i_p-e_g curves.

spaces of the grid and the amplification factor is relatively large because of the close spacing of the end turns of the control grid. As the bias is increased, the electron flow is cut off through the narrow spaces at the ends of the grid structure. However, they are still able to pass through the relatively large spaces at the center of the grid. The increased bias causes a decrease in the amplification due to the coarser turns in the central portion of the grid. A much greater value of bias is required to cut off the plate-current flow in this type of tube. The remote-cutoff tube is so named because the cutoff bias value is greater than (more remote from) the value required to cut off plate-current flow in tubes of evenly spaced turns.

Figure 2–18, B, shows the i_p-e_g curves for both a conventional sharp cutoff tube and a variable-mu or remote-cutoff tube. The cutoff bias for the normal tube is -5 volts, and because the slope is almost constant any change in bias produces little change in amplification. Contrasted with this characteristic, the curve for the variable-mu tube has a pronounced change in slope as the grid bias is increased from -10 volts to -15 volts and a small value of plate current is still flowing at a bias of -25 volts. The changing slope of this curve indicates a variation of amplification with bias. Thus, if a variable-mu tube is used with a bias source that varies with the signal strength, the output signal can be made substantially independent of the input signal strength. Automatic-gain-control circuits employing variable-mu tubes are discussed in connection with "Receivers" in chapter 12.

Multigrid Tubes

Electron tubes may be constructed with 4, 5, or 6 grids (fig. 2–19) in order to obtain certain characteristics. The grids may be used to influence the plate-current flow by introducing additional signal voltages having different frequencies, as in pentagrid converters or pentagrid mixers used in superheterodyne receivers. These applications are treated in chapter 11.

Multiunit Tubes

To reduce the number of tubes in radio circuits, the electrodes of two or more tubes frequently are placed within one envelope. Multiunit tubes generally are identified according to the way the individual types contained in the

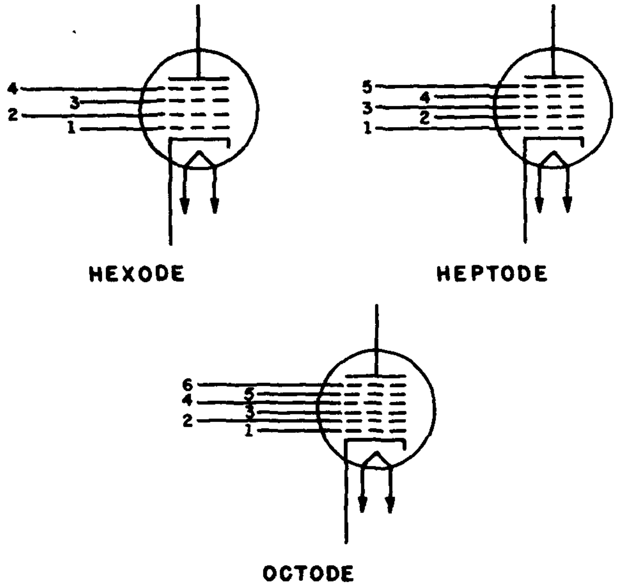


Figure 2–19.—Schematic diagrams of multigrid tubes.

envelope would be identified if they were made as separate units. Thus, a multiunit tube may be identified as a duplex-diode, a diode-pentode, a diode-triode-pentode, a pentagrid converter, and so forth. A number of multiunit tubes are shown in figure 2–20.

TUBES OPERATING AT ULTRAHIGH FREQUENCIES

As the operating frequency is increased, the capacitive reactance between electrodes in electron tubes decreases—that is,

$$X_C=\frac{1}{2\pi fC}.$$

At frequencies higher than 100 megacycles, the interelectrode capacitance of an ordinary electron tube provides a low-impedance path which shunts the external circuit. Also at these frequencies the electron transit time between cathode

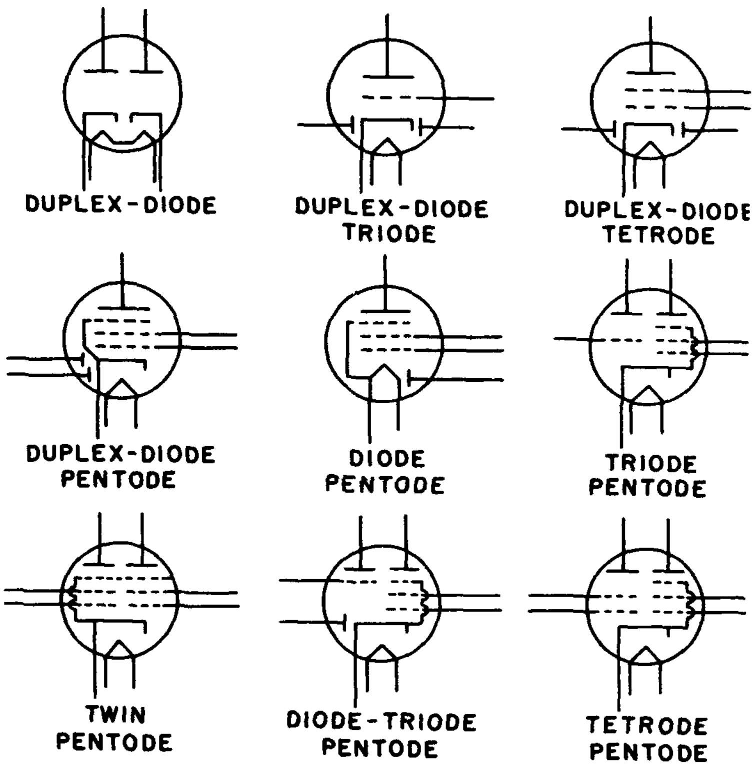


Figure 2-20.—Schematic diagrams of multiunit tubes.

and plate becomes appreciable. The transit time is about one-thousandth of a microsecond. As insignificant as this interval of time may seem, it nevertheless approaches and sometimes equals the time of one cycle of the applied signal and thus causes an undesirable shift in phase.

Ordinary Tubes

A small number of ordinary tubes can be operated at frequencies higher than 100 megacycles under certain critical operating conditions. The most suitable tubes of this type are triodes having low-interelectrode capacitance, close spacing of the electrodes to reduce transit time, a high-amplification factor, and a fairly low plate resistance. Because some of these requirements are conflicting, a compromise has to be made and tubes that strike a medium between the conflicting values are generally selected.

Special Ultrahigh-Frequency Tubes

The objectionable features of ordinary electron tubes are minimized considerably in the construction of special u-h-f tubes. These tubes have very small electrodes placed close together, and often have no socket base. By a proportionate reduction in all physical dimensions of a tube, the interelectrode capacitances are decreased without affecting the amplification factor or the transconductance. The electron transit time is likewise reduced.

ACORN electron tubes (fig. 2–21) have been developed especially for u-h-f operation and are available as diodes, triodes, and r-f pentodes. These tubes are very small physically and have closely spaced electrodes and no base. The tube connections are brought out to short wire pins sealed in the glass envelope. Such tubes are not used extensively because of limited power capabilities.

An enlarged version of the acorn tube, known as the DOORKNOB tube, can be operated at a considerably higher power level and at frequencies as high as 600 megacycles.

GAS-FILLED TUBES

In the manufacture of high-vacuum tubes, as much of the air as possible is removed from the envelope. In some cases low-vacuum tubes are designed purposely to contain a specific gas in place of air—usually nitrogen, neon, argon, or mercury vapor.

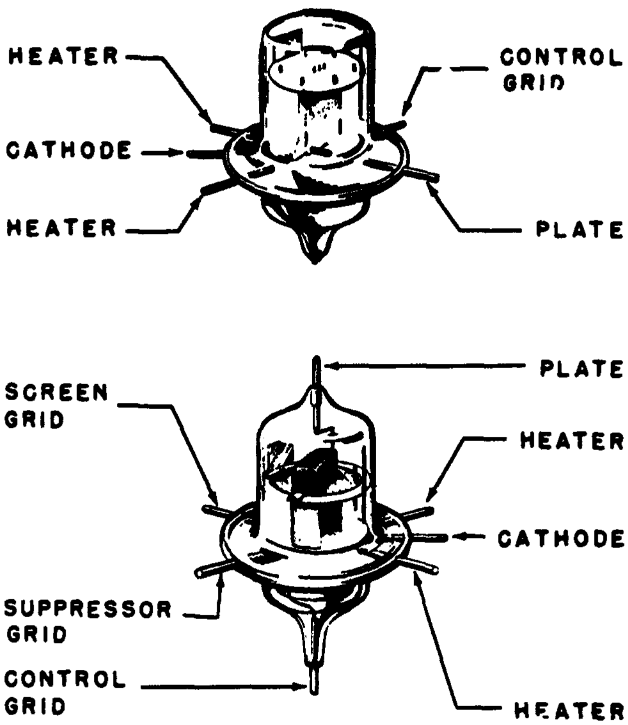


Figure 2-21.—Acorn electron tubes.

In a high-vacuum triode, the grid retains complete control of the current flowing in the tube. However, in a gas-filled tube the grid loses control when the tube ionizes, because of a sheath of positive ions that surround the grid. The plate current then rises rapidly to its full value. In this respect the gas tube acts like a snap-action switch. When the plate voltage falls below the deionization potential, the gas is deionized.

The gas-filled tube normally has a higher plate current rating than a high-vacuum tube of the same physical dimensions. When ionization occurs, the tube presents a lower impedance to the external circuit. Several types of gas-filled tubes are represented in figure 2–22. The small dot within the circle indicates that the tube is gas-filled.

Electrical Conduction in Gas Tubes

In a gas-filled tube, such as the diode of figure 2–22, A, the electron stream from the hot cathode encounters gas molecules on its way to the plate. When an electron collides with a gas molecule the energy transmitted by the collision

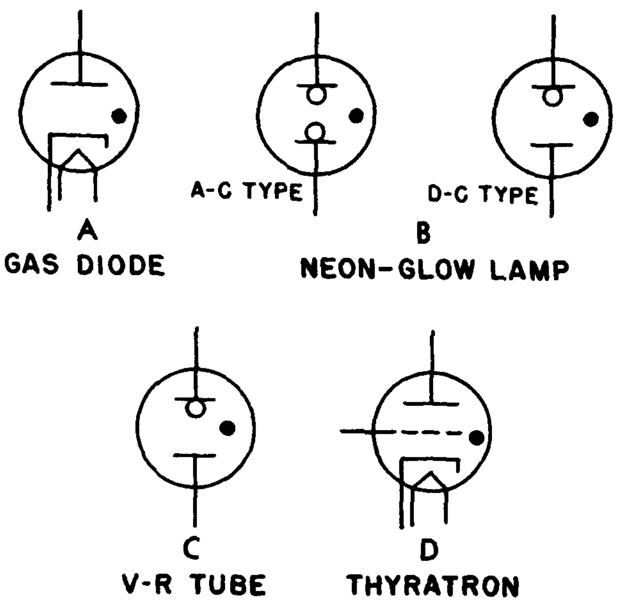


Figure 2–22.—Schematic diagrams of typical gas-filled tubes.

may cause the molecule to release an electron. This second electron may then join the original stream of electrons and thus be capable of liberating other electrons through collision with other gas molecules. This process which is cumulative is a form of ionization. The molecule that has lost an electron is called an ION and bears a positive charge. The tube in its ionized condition contains molecules, ions, and free electrons within the envelope. The positive gas ions are relatively large and in the vicinity of the cathode they neutralize a portion of the space charge. Thus electrons

flow from cathode to plate with less opposition than in a high-vacuum tube.

The heavier positive ions are attracted toward the negative cathode and while moving toward it they attract additional electrons from the space charge.

The energy needed to dislodge electrons from their atomic orbits and to produce the ionization is supplied by the source which supplies the voltage between the plate and cathode. There is a certain voltage value for a particular gas-filled tube at which ionization begins. When ionization occurs large currents flow at relatively low voltage across the tube. The voltage at which ionization commences is known as IONIZATION POTENTIAL, STRIKING POTENTIAL, or FIRING POINT.

After ionization has started, the action maintains itself at a voltage considerably lower than the firing point. However, a minimum voltage is needed to maintain ionization. If the voltage across the tube falls below this minimum value, the gas deionizes and conduction stops. The voltage at which current ceases to flow is known as the DEIONIZING POTENTIAL or the EXTINCTION POTENTIAL. The tube may therefore be used as an electronic switch that closes at a certain voltage and permits current to flow and then opens at some lower voltage and thus blocks the flow of current. Such a tube has almost infinite resistance before ionization and very low resistance after ionization.

Limitations in the Use of Gas Tubes

One limitation in the use of gas in electron tubes is the possibility that the tube will permit current to flow in the reverse direction (ARCBACK) when the plate has a high negative (inverse) voltage with respect to the cathode. The peak inverse voltage rating varies inversely with the temperature and pressure of the gas.

A second limitation is the possibility that the cathode may be destroyed by positive-ion bombardment as the plate voltage is increased to a high value. Because the mass of

the ion is very much greater than that of the electron, the result of its impact on the cathode may be serious, especially if the cathode is of the oxide-coated type. If the plate voltage is raised to a sufficiently high value, double ionization (two electrons dislodged from the gas molecule) occurs and the resultant increased velocity of the ions caused by the increased positive charge may quickly damage the emitter surface. The solution is to keep the plate voltage below the double ionization potential or to use a more rugged emitter which unfortunately will also have a higher work function.

Another limitation at high-operating frequencies is the possibility that arcback will occur because too many ions remain between the plate and cathode on the negative half cycle. This condition results from the fact that at high frequencies there is not sufficient time for the ions to be neutralized by the electrons before the full reverse voltage is applied. Because arcback causes the tube to offer a low resistance on both halves of the cycle, the power dissipated is increased and the tube will probably be destroyed. At high operating frequencies arcback may occur at a fairly low voltage and hence the tube is said to have a low INVERSE VOLTAGE RATING.

Gas Diodes

The neon-glow lamp or neon bulb (fig. 2–22, B) is a cold-cathode gas-filled diode. The cathode may have the same shape and size as the plate so that the tube can conduct in either direction depending only on the applied potential, or the structures of the cathode and plate may be such as to permit conduction in only one direction (fig. 2–22, C). Because the cathode is not heated in this type of tube no electrons are emitted to help in the ionization process. Therefore the firing potential for a neon-glow tube is higher than that for a tube in which a hot cathode is used, and the neon-glow tube is somewhat erratic in that the firing potential varies during operation. The passage of current through the tube is indicated by a glow whose color depends on the

gases that may be mixed with the neon. The glow is on the negative electrode or cathode. When an alternating voltage is applied both electrodes are alternately surrounded with a glow discharge.

A neon-glow tube placed in an r-f field of sufficient strength to ionize the gas in the tube will indicate the presence of such a field by glowing. A glow tube may also be used as a voltage regulator (chapter 3). Additional uses of glow tubes are as a source of light, as a part of a relaxation oscillator, as a rectifier, and to control circuit continuity in noise limiters.

Hot cathode, mercury-vapor diodes are specially designed to serve as rectifiers. Tubes of this type can pass much higher currents than high-vacuum tubes because the ionization of the mercury vapor partially dispels the cathode space charge. Mercury vapor is formed in these tubes when the small amount of liquid mercury enclosed in the envelope is vaporized by the hot cathode. These tubes are not capable of supplying their rated output until the mercury is completely vaporized. The relatively high voltage existing between the plate and cathode, before the tube begins to conduct load current, causes a large increase in the electron velocity. These high-velocity electrons cause the gas ions to acquire a higher positive charge and thus to bombard the cathode with a greater impact that is high enough to disintegrate the emitter surface if the action is allowed to continue for even a short period of time. Therefore sufficient time must be allowed for the tube to become heated before the plate voltage is applied.

Thyratrons

A gas-filled triode (fig. 2–22, D) or tetrode in which a grid is used to control the firing potential is called a THYRATRON. The grid in this tube functions somewhat the same as that of an ordinary electron tube, but the resultant control action is entirely different. Figure 2–23 shows the grid control characteristics of a typical thyratron.

Thus at a given plate voltage, for example 800 volts, the bias would have to be reduced from −10 volts to −8 volts

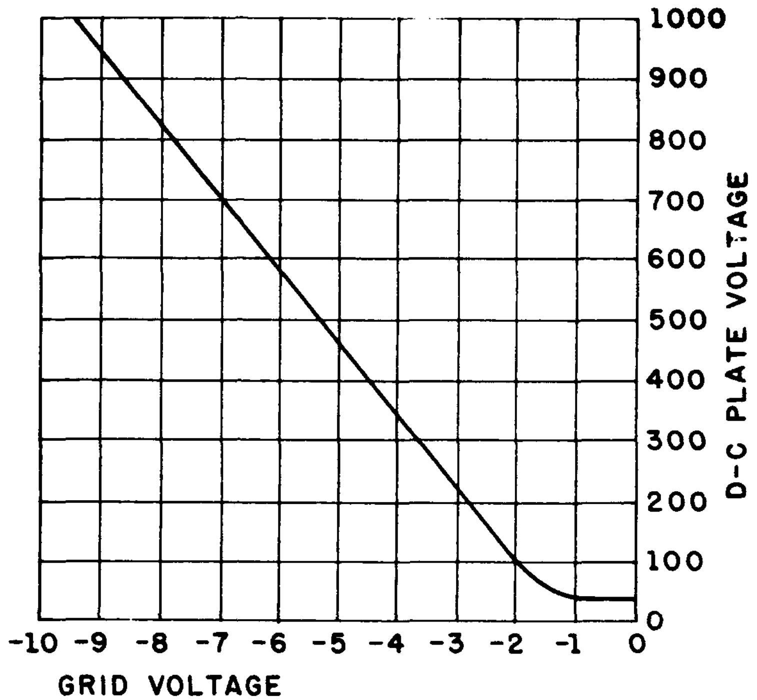


Figure 2-23.—Grid control characteristic of a typical thyratron.

before the tube would begin to conduct. Likewise, at a plate voltage of 300 volts the tube would begin to conduct at a grid potential of approximately −4 volts. When conduction starts, the grid loses control over the plate current and is no longer effective as a control element. To stop plate current flow, the plate voltage must be reduced below the ionizing potential. The grid operates in this manner because when conduction starts, positive ions are formed as a result of collisions and some of these ions are attracted to the negative grid. A positive-ion sheath is formed around the grid, thus destroying its effectiveness as a control element. Other positive ions move toward the cathode and neutralize the space charge. These two actions account for the fact

that once current flow starts, the grid loses control and the current rises rapidly to a large value.

Thyratrons have many practical applications in relay and trigger circuits.

CATHODE-RAY TUBES

Cathode-ray tubes are electron tubes of a special construction that permit the visual observation of current and voltage waveforms. A discussion of their construction and operation is included in chapter 13.

Electron-Ray Tubes

The electron-ray tube, or MAGIC EYE, contains two sets of elements, one of which is a triode amplifier and the other a cathode-ray indicator. The plate of the triode section is internally connected to the ray-control electrode (fig. 2–24, A) so that as plate voltage varies with the applied signal,

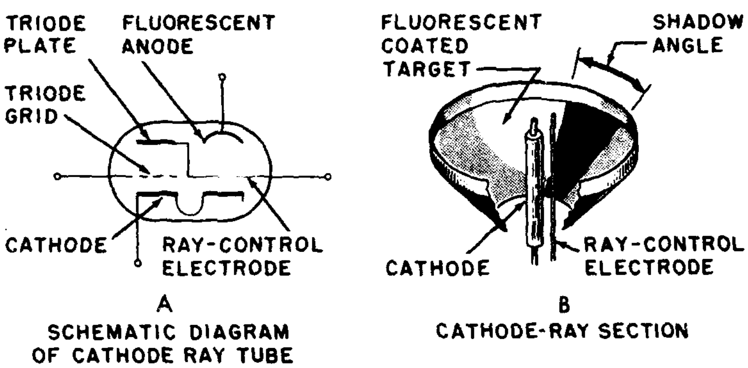


Figure 2–24.—Electron-ray tube.

the voltage on the ray-control electrode also varies. The ray-control electrode is a flat, metal strip so placed relative to the cathode that it deflects some of the electrons emitted from the cathode. The electrons that strike the anode, or target, cause it to fluoresce, or give off light. The deflection

caused by the ray-control electrode prevents electrons from striking part of the target; thus a wedge-shaped shadow is produced on the target. The size of this shadow is determined by the voltage on the ray-control electrode. When this electrode is at approximately the same potential as the fluorescent anode, the shadow disappears.

If the ray-control electrode is less positive than the anode, a shadow appears, the width of which is dependent upon the voltage on the ray-control electrode. If the tube is calibrated, it may be used as a voltmeter when rough measurements will suffice. However, the principal uses of the magic-eye tube are as a tuning indicator, in receiving sets and as a balance indicator in bridge circuits.

ANALYSIS OF TUBE ACTION.—The width of the shadow angle depends upon the relation of the voltage between the ray-control electrode and ground compared to the voltage between a point on the electric field gradient and ground in the vicinity of the ray-control electrode, as indicated in figure 2–25, A.

With no signal applied to the grid of the triode section, plate current is 240 μa (fig. 2–25, B). The voltage on the ray-control electrode is equal to the plate supply voltage less the drop through the 1-megohm resistor, or $250-240=10$ volts. The electric field gradient is assumed to vary as a straight line starting at the cathode with zero potential and terminating at the anode with a potential of $+250$ volts with respect to the cathode. A point on the electric field gradient in the vicinity of the ray-control electrode has a potential of $+50$ volts with respect to ground. Thus the ray-control electrode is negative with respect to the field at this point by an amount equal to $-(50-10)$, or -40 volts. The negative charge repels electrons and the shadow angle is established.

In figure 2–25, C, a 5-volt signal is developed between grid and ground of the triode section of the magic-eye tube. The plate current is reduced to 200 μa and the potential of the ray-control electrode is equal to $250-200$, or $+50$ volts

with respect to ground. Since the potential of a point on the electric field gradient in the immediate vicinity of the ray-control electrode is also +50 volts with respect to ground,

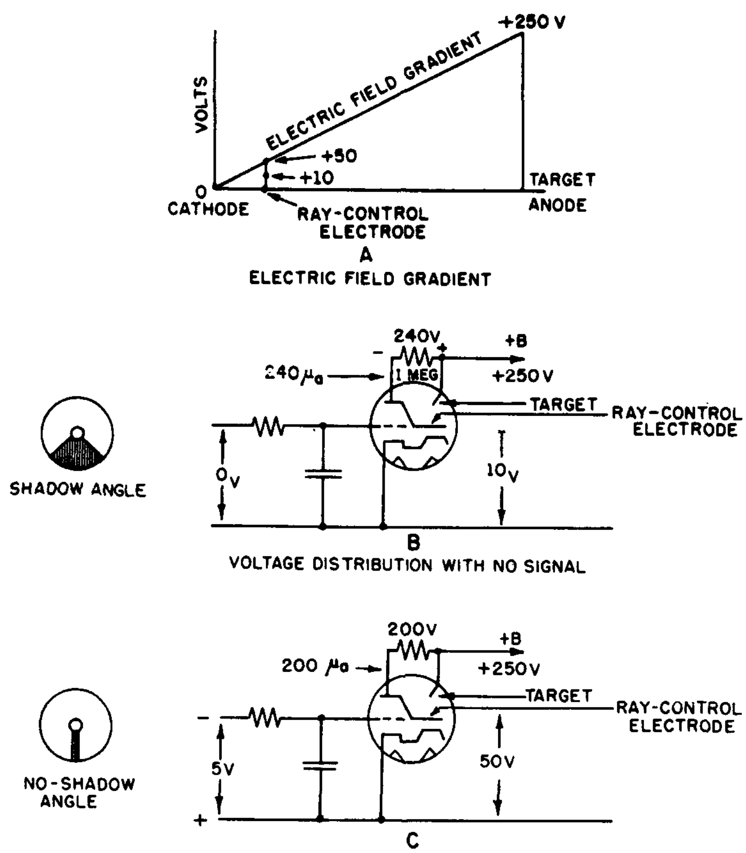


Figure 2-25.—Analysis of magic-eye tube action.

there is no difference in potential between the control electrode and the field. Thus, the control electrode does not repel electrons and the shadow angle closes, indicating the signal voltage applied to the triode grid.

QUIZ

1. What are the three principal uses of electron tubes?
2. In thermionic emission, what causes the electrons to gain enough energy to escape from the emitter?
3. In photoelectric emission, what determines the velocity of the emitted electrons?
4. What is the principal advantage of tungsten emitters?
5. In what classes of tubes are thoriated-tungsten emitters used?
6. What type of emitter is used in most types of receiving tubes?
7. Why are indirectly heated cathodes seldom used in portable equipment?
8. Why are electron tubes generally evacuated?
9. In electron-tube operation what is the meaning of the term SATURATION VOLTAGE?
10. At low values of plate voltage, how is plate current controlled?
11. In a tube employing what type of emitter is it unlikely that the plate current will ever be entirely independent of the plate voltage?
12. What is the basic use of diodes?
13. What is the term used to describe the smallest negative grid voltage (with respect to the cathode) that will stop the flow of plate current?
14. When is power consumed in the grid circuit?
15. What distinguishes dynamic from static characteristics?
16. Which of the tube characteristics gives a comparison between the relative effect of changes in plate and grid voltage on the plate current?
17. Which of the tube characteristics is determined when the plate voltage is varied and the grid voltage is held constant?
18. Give the expression for mu in terms of r_p and g_m.
19. Which of the tube characteristics indicates directly the ratio of the plate current change to the initiating grid voltage change?
20. What is the effect of operating an amplifier tube on the nonlinear portion of the i_p-e_g characteristic curve?
21. How are interelectrode capacitances of tubes reduced?
22. What advantages do tetrodes have over triodes?
23. What is the action called when plate current in a tetrode decreases as plate voltage increases?
24. What is the purpose of the suppressor grid in a pentode?
25. In a beam-power tube, what action prevents electrons emitted from the plate from reaching the screen grid?
26. What type of tubes is generally employed in the r-f section of radio receivers employing automatic volume control?
27. What two factors limit the usefulness of ordinary tubes when operated at ultrahigh frequencies?

28. Normally, how are gas-filled tubes cut off after the grid has lost control of the plate-current flow?
29. What are the relative plate current and internal impedance of gas-filled tubes compared with high vacuum types of the same physical dimensions?
30. From what source (plate supply or filament supply) does the energy come that causes the ionization in gas-filled tubes?
31. Name three limitations in the use of gas-filled tubes.
32. Why must the filaments of hot-cathode mercury-vapor diodes used as rectifiers be turned on for a definite period of time before the plate voltage is applied?
33. How does the action of the control grid in a thyratron differ from the grid action of a high-vacuum triode?
34. What is the principal use of the electron-ray (magic-eye) tube?

CHAPTER 3

POWER SUPPLIES FOR ELECTRONIC EQUIPMENTS

INTRODUCTION

The electron tubes in the various electronic equipments used in the Navy require voltages of various values for their filament, grid, screen, and plate circuits. It is the function of a power supply to supply these voltages at the necessary current ratings. Except for filament power, which may be alternating current, the output from a power supply must be nearly pure direct current, and the voltage must be of the correct value for the circuits of the equipment being used. Transmitters require more power than receivers, and consequently transmitter power sources must supply higher voltages with greater current ratings than receiver power sources.

Power used to heat the filaments of tubes is sometimes called the A SUPPLY and is normally furnished at a low voltage, with a relatively high current drain. In portable and mobile equipments filament power is furnished by batteries, generators, or dynamotors. In permanent land-based or shipboard installations filament power is obtained from the standard a-c line via a step-down transformer.

The plate and screen power supply, normally called the B SUPPLY, furnishes power at a relatively high voltage and low current. In portable equipment the plate voltage may

be supplied by batteries or a hand generator. Mobile sets may employ dynamotors or vibrators operated from batteries to generate the high voltage necessary for the plates and screens. Permanent installations ordinarily use a transformer-rectifier-filter system for the high-voltage plate and screen supply.

When a separate grid-bias voltage is used, it is sometimes called the C SUPPLY. In many instances grid-bias voltage is obtained by some means of self-bias. However, for large power-amplifier tubes, or where self-bias is not satisfactory, a separate transformer-rectifier-filter system or a d-c generator may be employed.

Radio power supplies may be conveniently divided into three general classes—battery, alternating current, and electromechanical systems. However, in this chapter primary consideration is given to obtaining the necessary d-c potentials from an a-c source via a transformer-rectifier-filter system. Dynamotors and vibrator power supplies are also treated briefly in this chapter; but batteries and generators are included in basic electricity texts and hence are not discussed in this chapter.

CATHODE HEATING POWER

Directly Heated Cathodes

When alternating current is applied to the directly heated cathode, careful consideration must be given to the possible effects of the reversing polarity of the heater voltage upon the output signal. Regardless of the measures taken to prevent it, a component of the heating current will modulate the space current of the tube. This component is called HUM because it produces a humming sound in the audio output of an equipment.

In an effort to minimize this hum the following circuit arrangements (fig. 3–1) are utilized: (1) center-tapped resistor, (2) center-tapped filament transformer with separate bias, and (3) center-tapped transformer with cathode bias.

It is important to return the grid and plate circuits of

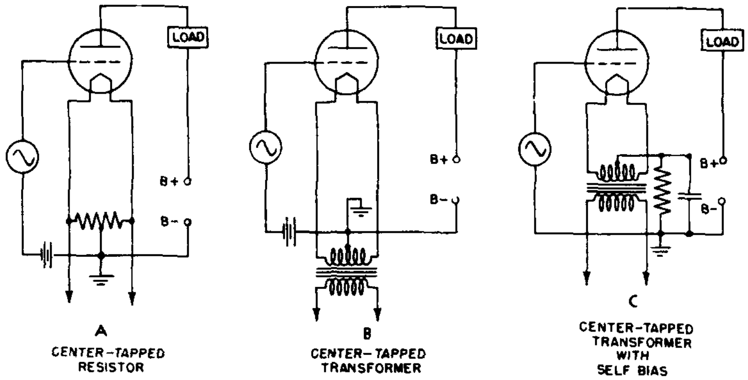


Figure 3-1.—Filament connections for reducing hum.

a tube to a point in the exact electrical center of the filament circuit, as shown in figure 3-1, A. In this figure the resistance of the resistor is high with respect to the resistance of the filament, so that most of the current flows through the filament. Tapping the resistor at its electrical center is essentially the same as tapping the filament at its electrical center.

In figure 3-1, B, the grid return path may be traced through the two halves of the center-tapped filament winding simultaneously outward from the center tap to the two ends of the winding. At any instant the a-c filament voltage makes the grid more negative with respect to one end of the filament and less negative with respect to the other. The increases and decreases from one end of the filament to the other tend to cancel each other and therefore the bias (cathode-to-grid potential) remains the same. In figure 3-1, C, the same action takes place except that cathode bias is developed across the resistor connected between center tap and ground.

The magnetic field set up by the current that passes through the heater wires may induce a power-frequency hum in other tube elements by electromagnetic induction. The effect of this coupling can be materially reduced by using

twisted filament leads. The current in each wire flows in opposite directions and the magnetic flux fields tend to cancel each other.

Even though the effects of the fundamental frequency of the heater current are essentially eliminated by center-tapping the filament, another serious problem presents itself—the output of the tube will contain an a-c component whose frequency is twice that of the fundamental. This condition results from the very nature of alternating current. The current reaches a peak value twice during each cycle—once during each direction of current flow. Therefore, twice during each cycle the filament is heated to maximum temperature and the maximum number of electrons is emitted. In order to minimize this effect, large heavy cathodes are used to provide enough thermal inertia to prevent the temperature of the filament from changing significantly with voltage alternations.

Indirectly Heated Cathodes

Tubes using indirectly heated cathodes do not require the center-tap return because the emitting element and the heating element are electrically insulated. However, the frequency of the filament voltage can be introduced into the electron stream by capacitive coupling. The effects of this coupling are minimized by operating the cathode and filament at or near the same potential. Capacitive coupling between the filament power leads and other external tube circuits is reduced by the proper placement of the twisted filament leads.

B-VOLTAGE SUPPLIES

The B-supply for Navy electronic equipment is generally obtained via a transformer-rectifier-filter system from the ship's a-c bus. The voltage and phase considerations are dependent upon the type of ship. On ships having direct current as the power source, a motor-generator is generally utilized to supply alternating current to the electronic equipment. There are many types of low-power emergency com-

munications equipment that use storage batteries or dry batteries for power sources. Such equipments use vibrator-type rectifier circuits or dynamotors to convert the low d-c voltage into a d-c voltage sufficiently high for the plates and screen grids.

Airborne equipments use dynamotors for the high-voltage B-supply. Inverters are also used in airborne equipments. An inverter utilizes a low-voltage d-c supply to operate a d-c motor which in turn drives an a-c generator. The a-c generator supplies power generally at 400 or 800 cycles. Electromechanical systems such as these are treated later in this chapter.

If it is assumed that a suitable a-c supply is available, the problems involved in obtaining a suitably high d-c potential are as follows: (1) If the voltage is not sufficiently high, a step-up transformer must be provided; (2) the voltage must be rectified—that is, changed into pulsating direct voltage; (3) the ripples must be removed; and (4) some form of voltage regulation must be employed. Most of the remainder of this chapter is devoted to these problems.

Rectifiers for Power Supplies

The majority of rectifier systems employing electron tubes utilize either high-vacuum or gas-filled tubes. The high-vacuum diode is the most widely used in low-current applications. The hot-cathode mercury-vapor tube is widely used in high-current applications. The presence of mercury vapor in the tube envelope reduces the vacuum and results in low internal resistance, thus allowing a large amount of current to be drawn.

On the other hand, dry-disk rectifiers, such as selenium and copper-oxide rectifiers, do not employ electron tubes. Selenium rectifiers are sometimes used as plate-supply rectifiers in small radio receivers. Copper-oxide rectifiers have miscellaneous applications, for example, they may be used as instrument rectifiers or bias-supply rectifiers. Both of these rectifiers have the advantage of not requiring heater current or warmup time.

HIGH-VACUUM RECTIFIERS.—A diode acts as a rectifier because it passes current in only one direction—that is, from cathode to plate when the plate is positive with respect to the cathode. The characteristics of diodes are discussed in chapter 2. The important characteristics of the high-vacuum rectifier tube are its maximum peak plate current and its maximum inverse peak plate voltage ratings.

The peak plate current is limited by the number of electrons emitted by the cathode and therefore is dependent upon cathode construction. It is evident that current in a rectifier never flows for more than one-half of each a-c cycle. At the power frequency, the d-c output current as indicated by a meter is less than one-half the peak plate current.

The peak inverse plate voltage is the peak negative voltage that is applied to the plate during the portion of the cycle when the tube is not conducting. The d-c output voltage and peak inverse voltage vary with the type of circuit. In general, the peak inverse voltage is equal to or twice the peak value of the d-c output voltage.

An important factor in circuit design utilizing high-vacuum tubes is the voltage drop across the tube during the conducting half cycle. In most high-vacuum rectifiers this voltage drop is relatively large compared to the drop in gas-filled tubes and is the limiting factor in the design of high-current electron-tube rectifiers. The drop increases with an increase in current and results in poor regulation for high-current applications. High-vacuum rectifiers are used almost universally in receiver power supplies and also in many high-voltage, low-current applications.

The types of rectifiers used in receivers are generally provided with directly heated, oxide-coated filaments. Most of these rectifiers consist of two units in a single envelope because they are generally used in pairs for full-wave operation. The FULL-WAVE RECTIFIER, as this type of tube is called, is provided with a single filament to supply both plates. Lower-power tubes with indirectly heated cathodes are available for special applications in which the cathode and heater supplies must be electrically isolated.

In addition to low-power applications, high-vacuum rectifiers for high-voltage applications have been designed to withstand a peak inverse voltage of 100,000 volts. Commercial units have been built to provide peak plate currents as high as 7.5 amperes. High-vacuum rectifiers are seldom used for high-voltage high-current applications except where inherent ruggedness outweighs the other disadvantages. The mercury-vapor rectifier is used in the majority of medium- and high-power equipments.

MERCURY-VAPOR RECTIFIERS.—The introduction of mercury-vapor tubes to the electronics industry was one of the greatest single contributions to the development of high-powered electronic equipment. As a help in understanding the importance of the mercury-vapor tube, consider the foremost disadvantage of high-vacuum rectifiers. As stated previously, the voltage drop across a high-vacuum tube varies with the load current; and when the current varies widely, the regulation is poor. The high-vacuum rectifier used on heavy loads has a relatively high loss and low efficiency. In some high-power applications a water cooling system is employed to carry the heat from the tube elements. The power loss in an electron tube is usually of the order of 15 percent of the input power to the rectifier. In comparison, the power loss in a mercury-vapor rectifier is only about 1.5 percent of the total input.

The greater efficiency of the mercury-vapor rectifier is a result of the low-voltage drop across the tube. In normal operation this voltage drop rarely exceeds 15 volts, even when the tube is operating at very high values of load current. The filament of the high-vacuum rectifier is surrounded by a space charge which acts as a shield to impede electron flow between cathode and plate.

In a mercury-vapor rectifier, a small amount of mercury is introduced into the tube envelope. Because of the low pressure within the tube, the mercury vaporizes completely as the unit reaches normal operating temperature. If a positive potential is applied to the anode of the tube, electrons are emitted from the filament and move toward the

anode. Because the tube envelope is filled with mercury vapor, collisions occur between the moving electrons and the atoms of mercury.

Each collision knocks an electron of a mercury atom away from the influence of the nucleus. The atom, now minus an electron, becomes a positive ion. This ion is then drawn toward the negative filament and is promptly neutralized by one of the electrons forming the space charge around that element. Ionization of mercury vapor occurs when the potential gradient between the plate and cathode is 10.4 volts. Any further increase in plate voltage will ionize more atoms—each in turn neutralizing a space electron—until the drop reaches 15 volts, and at that time the tube will no longer show a plate-voltage rise for a proportional rise in current.

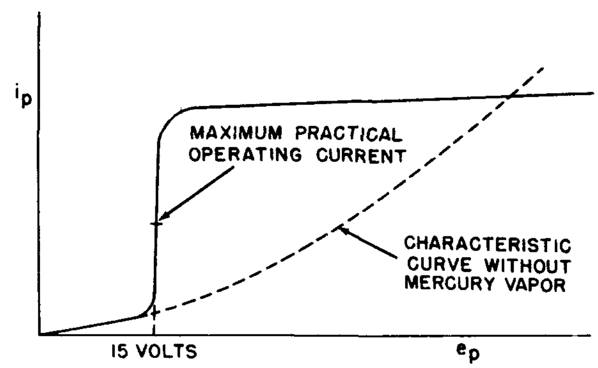


Figure 3-2.—Current vs voltage relation in a mercury-vapor diode.

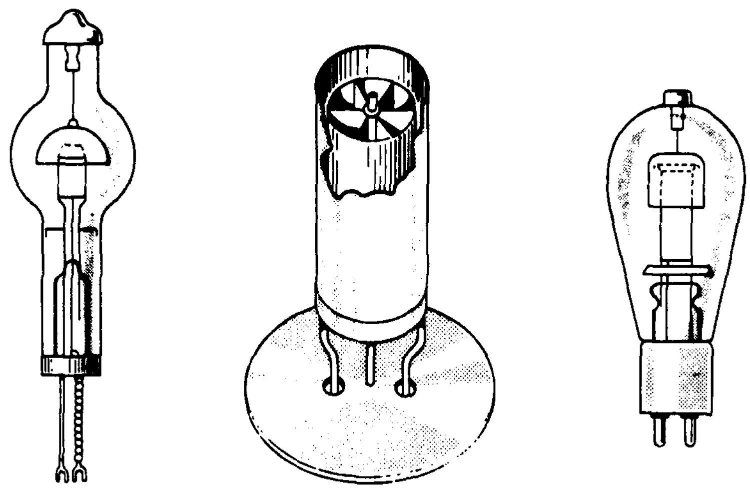
Figure 3-2 shows graphically the i_p-e_p relation during ionization. This figure indicates the most useful characteristics of a mercury-vapor tube—THE VOLTAGE DROP ACROSS THE TUBE REMAINS AT A CONSTANT VALUE OF 15 VOLTS REGARDLESS OF THE CURRENT FLOWING THROUGH THE TUBE, provided the rated tube current is not exceeded. On overload the voltage drop increases to some extent. When the tube drop exceeds 22 volts the filament may be damaged by excessive bombardment of positive ions. At potentials below 22 volts,

this bombardment is insufficient to cause damage to the filament.

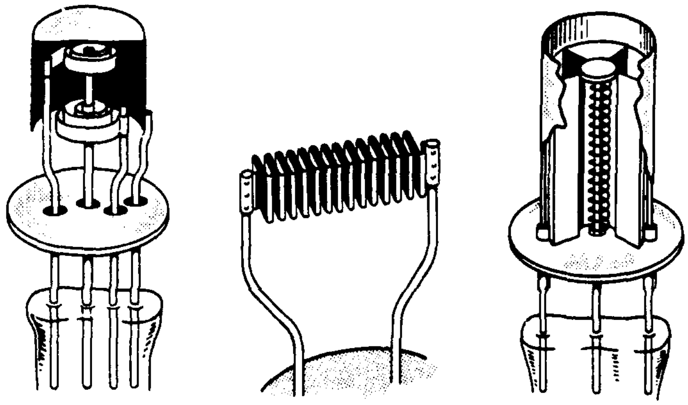
Another important characteristic of the mercury-vapor tube is its maximum inverse-voltage rating. This is the sparking voltage through the mercury vapor in a direction opposite to that of normal flow and is always less than it would be if the vapor were not present. A mercury-vapor tube always has a lower flashback voltage than a high-vacuum tube of similar construction. Nevertheless, mercury-vapor tubes with high inverse-voltage ratings have been developed. For example, a mercury-vapor diode having an output of 10 amperes and a maximum safe peak inverse voltage of 22,000 volts is used extensively in broadcast-transmitter power supplies.

In the practical operation of mercury-vapor rectifiers the vapor must reach its proper operating temperature before plate voltage is applied. If this precaution is not taken, the high voltage drop across the tube causes secondary emission from plate to cathode, and arcback occurs. In high-vacuum rectifiers the only factor considered in cathode heating is the emission of electrons. The cathode of the mercury-vapor tube must not only emit free electrons for conduction, but must also heat the surrounding space in order for the mercury-vapor temperature to be in the range of 20° to 60° centigrade. The cathode construction indicated in figure 3-3 facilitates this heating.

The heat given off by the inner turns of the spiral filament is absorbed by the outer turns. Radiation from the outer surface is reduced by a polished shield surrounding the filament. The plate (anode) is a metal cup fitting over the cathode. This arrangement reduces the tendency to arc back. It also shields the plate-cathode region from external electric fields. The graph shown in figure 3-4 plots flashback voltage and tube drop against operating temperature (C°). Below the indicated operating range of temperatures, the tube drop is excessive with inherent danger of positive-ion bombardment. At high operating temperatures, flashback potentials drop to an intolerably low value.



MERCURY-VAPOR TUBES WITH HEATER DETAILS



HOT-CATHODE EMITTING STRUCTURES

Figure 3-3.—Mercury-vapor diodes with hot-cathode emitting structure.

METALLIC RECTIFIERS.—When dissimilar metals are in contact, electrons flow in greater numbers in one direction across the area of contact than in the other direction.

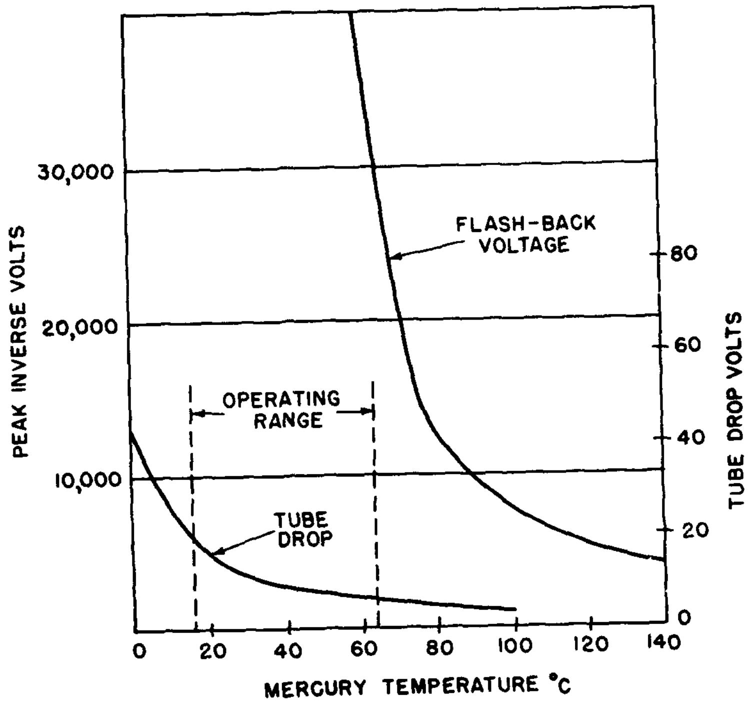


Figure 3-4.—Peak inverse voltage vs temperature.

Metallic rectifiers operate on this principle. The two combinations of substances that are most widely used for dry-disk rectifiers are: (1) A thin film of copper oxide (cuprous oxide) and copper, and (2) selenium and either iron or aluminum. Metallic rectifier units are represented by the symbol in figure 3-5, A. The arrowhead in the symbol points in the direction of the electron flow.

Figure 3-5, B, indicates how a metallic rectifier may be used in place of a diode rectifier. The waveforms indicate

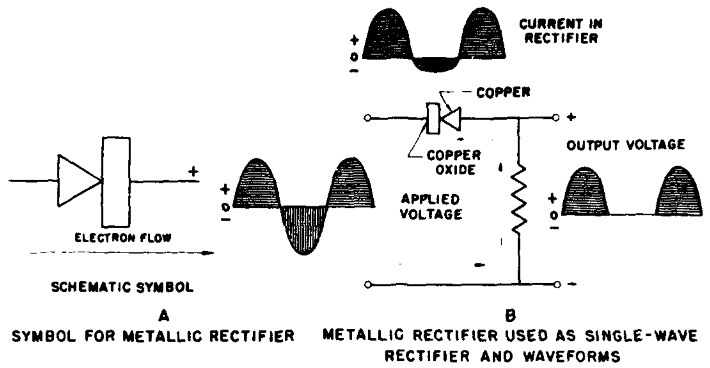


Figure 3-5.—Metallic rectifier symbol and waveforms.

that much more current flows in one direction than in the other. Although the copper-oxide rectifier is shown, the selenium rectifier may be used instead.

In the copper-oxide rectifier shown in figure 3-6, A, the

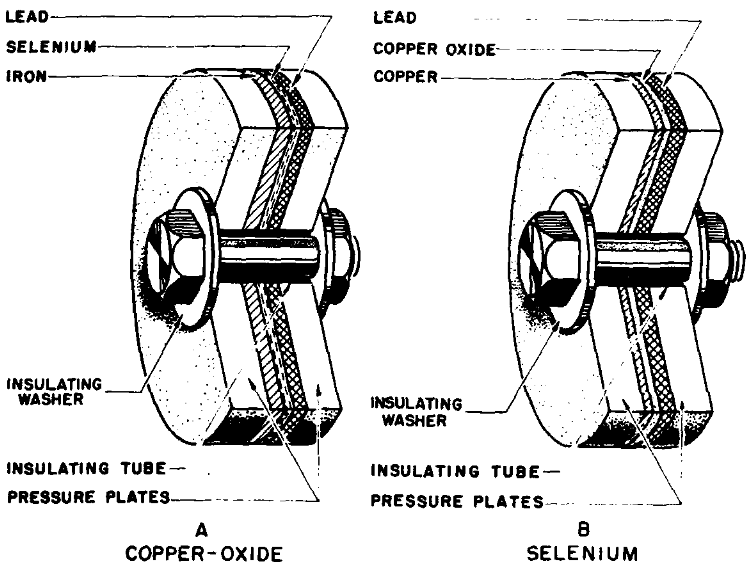


Figure 3-6.—Metallic rectifier construction.

oxide is formed on the copper by partial oxidation of the copper by means of a high temperature. In this type of rectifier the electrons flow more readily from the copper to the oxide than from the oxide to the copper. External electrical connection may be made by connecting terminal lugs between the left pressure plate and the copper and between the right pressure plate and the lead washer.

For the rectifier to function properly, the oxide coating must be very thin. Thus, each individual unit can stand only a low inverse voltage. Rectifiers designed for moderate and high-power applications consist of many of these individual units mounted in series on a single support. The lead washer enables uniform pressure to be applied to the units so that the internal resistance may be reduced. When the units are connected in series, they normally present a relatively high resistance to the current flow. The resultant heat developed in the resistance must be removed if the rectifier is to operate satisfactorily. Many commercial rectifiers have copper fins between each unit for the purpose of dissipating the excess heat. The useful life of the unit is extended by keeping the temperature low (below 140° F.). The efficiency of this type of rectifier is generally between 60 and 70 percent.

Selenium rectifiers function in much the same manner as copper-oxide rectifiers. A selenium rectifier is shown in figure 3–6, B. Such a rectifier is made up of an iron disk that is coated with a thin layer of selenium. In this type of rectifier the electrons flow more easily from the selenium to the iron than from the iron to the selenium.

Commercial selenium rectifier units are designed to pass 50 milliamperes per square centimeter of plate area. This type of rectifier may be operated at a somewhat higher temperature than a copper-oxide rectifier of similar rating. The efficiency is between 65 and 85 percent, depending on the circuit and the loading. As in the case of the copper-oxide rectifier, any practical number of units may be bolted together in series to increase the voltage rating. Larger element disks and the necessary cooling fins may be used

for higher current ratings. Also, forced-air cooling may be used.

Metallic rectifiers may be used not only as half-wave rectifiers, as shown in figure 3-5, but also in full-wave and bridge circuits. In each of these applications the action of the metallic rectifier is similar to that of a diode.

Metallic rectifiers may be used in battery chargers, instrument rectifiers, and many other applications including welding and electroplating. Commercial radios also frequently use selenium rectifiers in the high-voltage power supply, as do other electronic equipments.

Rectifier Circuits

HALF-WAVE RECTIFIER.—A half-wave rectifier is a device by means of which alternating current is changed into pulsating direct current by permitting current to flow through the device only during one-half of each cycle.

In a diode, electrons are attracted to the plate when it is more positive than the cathode. When the plate becomes negative with respect to the cathode, electrons are repelled by it and no electron stream can flow in the tube. Therefore, a single diode may be used as a half-wave rectifier because electrons can flow in the tube during only the half of the cycle when the plate is positive relative to the cathode.

Figure 3-7, A, shows a simple half-wave rectifier circuit. The primary winding of the transformer is shown connected to an a-c input source. The principal action of the transformer is to increase the voltage from 115 volts to a higher value in the main secondary winding. The action of the small secondary winding at the top of figure 3-7, A, steps down the 115 volts to a suitably lower voltage. It supplies heater current to the filament of the rectifier tube. The actual connections of this winding are not shown in the figure, but are indicated by the symbol *XX*.

The upper end of the high-voltage secondary winding is connected to the plate of the diode and the other end is connected to a junction with ground and a load resistance

represented by the resistor, R. The load resistor is connected to the cathode and is in series with the tube.

The operation of the circuit is illustrated by the waveforms at the right of figure 3–7. The alternations of the input

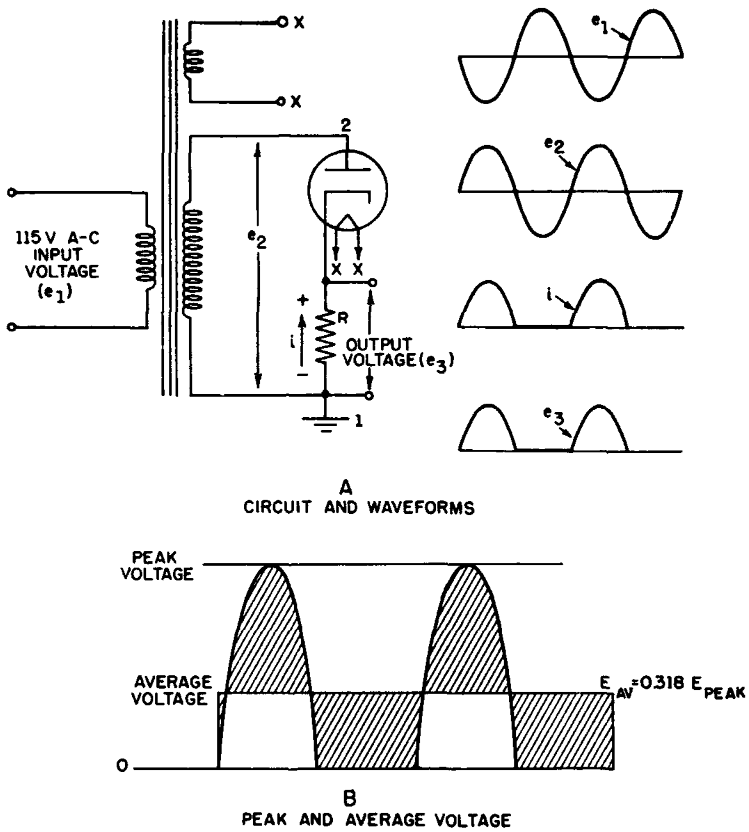


Figure 3–7.—Simple half-wave rectifier circuit and waveforms.

voltage, e_1, are reproduced by the transformer with an increase in voltage, e_2, in the secondary winding. The waveforms indicate a 180° difference in phase between e_1 and e_2. This difference is characteristic of induced voltages. The induced secondary voltage, e_2, is impressed across the diode

and its series load resistance. This voltage causes current i to flow through the diode and its series resistor on the positive half cycles (when the plate is positive). The resultant output voltage, e_3, across the load has a pulsating waveform, as shown in the figure. This pulsating waveform is called a RIPPLE VOLTAGE.

When point 2 is positive and point 1 is negative, electrons flow from the ground junction (1), through the load resistor (R), to the cathode, to the plate, and thus to the upper terminal (2) of the transformer. The secondary winding thus acts as the immediate source of voltage for the current flow.

Because current flows in only one direction (point 1 to point 2) through the diode and its load, the polarity of the load resistance is always as shown. The fact that the positive point of the load resistance is connected to the cathode of the tube may be confusing when the cathode is thought of as being the negative element of the tube. However, the only requirement necessary for conduction is that the plate be more positive than the cathode. The cathode can be positive with respect to other points in the circuit. The voltage drop across the tube is usually quite small when compared with that across the load resistance.

The load is connected to the cathode as shown rather than in the plate circuit to enable the use of a common ground for the negative side of the load and transformer winding, with ground continuity maintained throughout the entire input cycle.

The half-wave rectifier utilizes the transformer during only one-half of the cycle, and therefore for a given size of transformer less power can be developed in the load than could be developed if the transformer were utilized on both halves of the cycle. In other words, if any considerable amount of power is to be developed in the load the half-wave transformer must be relatively large compared with what it would have to be if both halves of the cycle were utilized. This disadvantage limits the use of the half-wave rectifier to applications that require a very small current drain. The

half-wave rectifier is widely used for commercial a-c d-c radio receivers and for the accelerating voltage supplies of oscilloscopes.

For a half-wave rectifier (assuming half sine waves, as shown) the rms voltage, E, in the tube is

$$E=\frac{e_{max}\times 0.707}{2},$$

and the average voltage (fig. 3–7, B) is

$$E_{av}=\frac{e_{max}\times 0.636}{2}.$$

The rms and average output currents are determined in the same way.

Because the d-c load current flows through the transformer secondary in only one direction, there is a tendency for the molecules in the iron core to become oriented in one direction. This effect is called D-C CORE SATURATION and reduces the effective inductance of the transformer. The net effective inductance with the d-c core saturation effect present is known as TRANSFORMER INCREMENTAL INDUCTANCE. Thus, the transformer incremental inductance is reduced with increasing d-c load current. The resultant effect is to decrease the primary counter emf to a greater degree and thus increase the load component of primary current correspondingly. Therefore, the efficiency of the transformer is reduced and the regulation is impaired. The output is far from being a continuous d-c voltage and current and for these reasons the half-wave rectifier circuit is seldom used for high current loads.

FULL-WAVE RECTIFIER.—A full-wave rectifier is a device that has two or more elements so arranged that the current output flows in the same direction during each half-cycle of the a-c supply.

Full-wave rectification may be accomplished by using two diodes in the same envelope (a dual diode) with a common

cathode connected to one end of the load resistance, as shown in figure 3-8, A. The other end of the load resistor is connected to the center tap, *C*, of the transformer secondary.

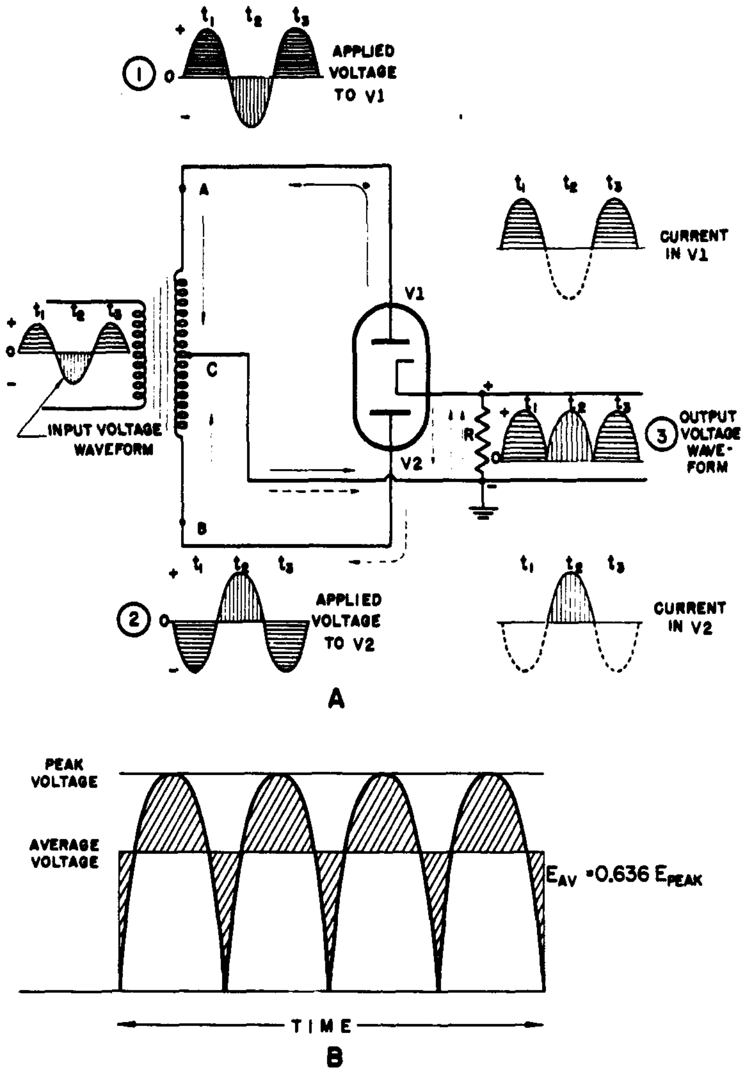


Figure 3-8.—Simple full-wave circuit and waveforms.

The two halves of the secondary winding AC and BC may be a center-tapped winding as shown, or they may be separate windings. In either case the load circuit is returned to a point midway in potential between A and B so that the load current is divided equally between the two tubes.

That part of the secondary winding between A and C may be considered a voltage source that produces a voltage of the shape shown at ① in figure 3–8, A. This voltage is impressed across the plate to cathode of $V1$ in series with load resistor R. During the half-cycle, marked t_1, the plate of $V1$ is positive with respect to its cathode. Therefore, electrons flow in the direction indicated by the solid arrows. This flow of electrons from ground up through load resistor R makes the cathode positive with respect to ground. Thus the load voltage is developed across R between cathode and ground. During this same half-cycle the voltage across BC is negative, as shown at ② in figure 3–8, A, and the plate of $V2$ is negative with respect to the cathode. Thus during the time $V1$ conducts, $V2$ is nonconducting. A half-cycle later, during interval t_2, the polarity of the voltages on the plates of the two tubes is reversed. $V2$ now conducts, and $V1$ is nonconducting. The electron flow through $V2$ is in the direction indicated by the dotted arrows. This current also flows from ground up through R and makes the cathode positive with respect to ground. Thus another half-cycle of load voltage is developed across R. A study of the figure shows that only one section of the twin diode is conducting at any given instant.

Because there are two pulsations of current in the output for each cycle of the applied alternating voltage, the full-wave rectifier utilizes the power-supply transformer for a greater percentage of the input cycle. Hence, the full-wave rectifier is more efficient than the half-wave rectifier, has less ripple effect, and may be used for a much wider variety of applications.

For a full-wave rectifier (having sine waves, as shown) the rms output voltage, E, across the load is

$$E=e_{max}\times 0.707,$$

and the average voltage (shown in fig. 3–8, B) is

$$E_{av} = e_{max} \times 0.636.$$

The rms and average output currents are similarly determined. In the case of both the half-wave and the full-wave rectifier no filtering is assumed in computing the rms and average values.

The d-c load current flows equally through the two halves of the transformer secondary and in opposite directions. Thus the ampere turns are equal and act in opposite directions so that there is no tendency to orient the molecules of the iron core in any one direction. Therefore the transformer inductance is not reduced as it is in the half-wave rectifier and both the voltage regulation and efficiency are improved. The full-wave rectifier is used widely in radio transmitters and receivers.

Bridge rectifier.—If four rectifiers are connected as shown in figure 3–9, A, the circuit is called a bridge rectifier. The input (waveform ①) to such a circuit is applied to diagonally opposite corners of the network, and the output is taken from the remaining two corners.

During one-half cycle of the applied alternating voltage point A becomes positive with respect to point B by the amount of the voltage induced in the secondary of the transformer. During this time, the voltage across AB may be considered to be impressed across a load consisting of $V1$, load resistor R, and $V3$ in series. The voltage applied across these tubes makes their plates more positive than their cathodes, and electrons flow in the path indicated by solid arrows. The waveform of this current is shown at ② and ③. One-half cycle later, $V1$ and $V3$ are nonconducting, and an electron stream flows through $V4$, R, and $V2$ in the direction indicated by the dotted arrows. The waveform of this current is shown at ④ and ⑤. The current through the load, R, is always in the same direction. As this current flows through R it developes a voltage having the waveform shown at ⑥. The bridge rectifier is a full-wave rectifier because current flows in the load during both halves of each cycle of applied alternating voltage.

One advantage of a bridge rectifier over a conventional full-wave rectifier is that with a given transformer the bridge circuit produces a voltage output nearly twice that of the full-wave circuit. This increase in voltage may be illustrated by assigning values to some of the components in figure 3-8, A, and 3-9. Assume that the same transformer is used in

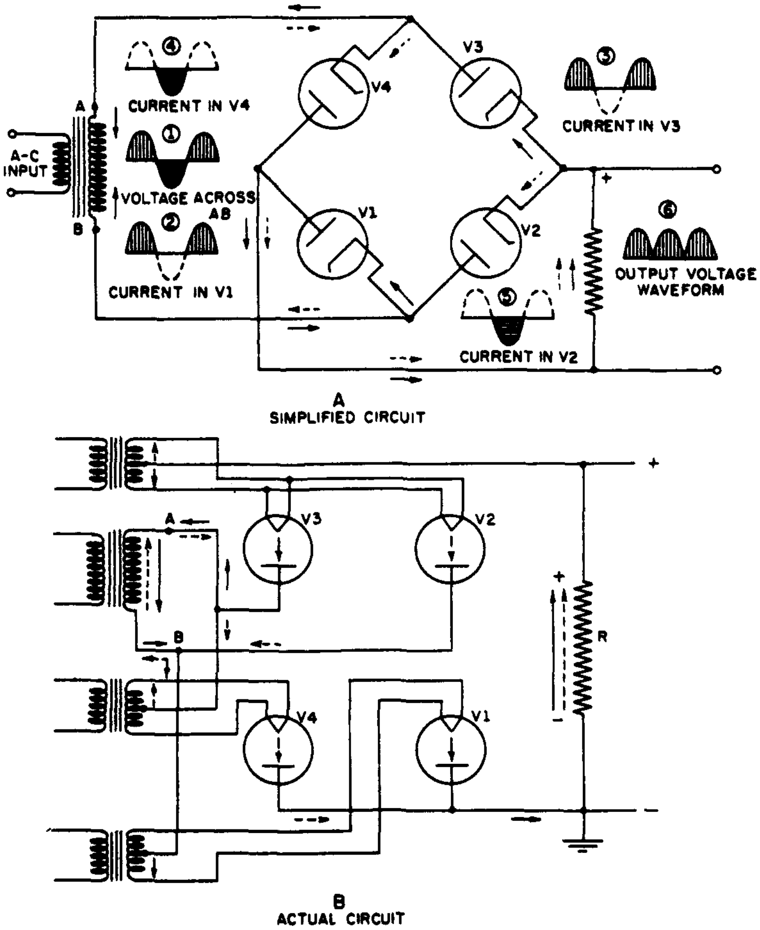


Figure 3-9.—Bridge-rectifier circuit.

both figures. Note that the center tap is not used in the bridge circuit. The peak voltage developed between A and B is assumed to be 1,000 volts in both figures. In the full-wave circuit in figure 3-8, A, the peak voltage from the center tap, C, to either A or B is 500 volts. Because only $V1$ or $V2$ is conducting at any instant, the maximum voltage that can be rectified at any instant is 500 volts. Therefore, the maximum voltage that can be developed across the load resistor, R, is 500 volts, less the small voltage drop across the tube that is conducting. In the bridge circuit of figure 3-9, however, the maximum voltage that can be rectified is the full voltage of the secondary of the transformer, or in this case 1,000 volts. Therefore, the voltage that can be developed across the load resistor, R, is 1,000 volts less the voltage drop across the two tubes that are conducting. Thus, the full-wave bridge circuit produces a higher output voltage than the conventional full-wave rectifier does with the same transformer.

A second advantage of the bridge circuit is that the peak inverse voltage across a tube is only half the peak inverse voltage impressed on a tube in a conventional full-wave rectifier that is designed for the same output voltage. For example, if the two circuits are to produce the same output voltage, the transformer secondary in the full-wave rectifier (fig. 3-8, A) has a 2,000-volt peak developed across it, while that for the bridge rectifier (fig. 3-9) has only a 1,000-volt peak. When $V1$ in figure 3-8, A, is not conducting, its plate is made negative relative to its cathode by a maximum voltage of 2,000 volts. The same is true for $V2$. This negative voltage is called the PEAK INVERSE VOLTAGE, which is a stress that tends to cause breakdown within the tube. In figure 3-9, however, when tubes $V1$ and $V3$ or $V2$ and $V4$ are not conducting, the peak inverse voltage for any one tube is then 1,000 volts, which is half of the peak inverse voltage across either tube in figure 3-8, A.

The bridge rectifier circuit has a disadvantage in that three filament transformers are required for the tubes. In the actual circuit in figure 3-9, B, the filament transformer

connections of $V2$ and $V3$ are operated at the same relative potential and therefore may be connected to the same filament winding.

The filaments of $V1$ and $V4$, however, are returned to opposite ends of the high-voltage secondary of the power-supply transformer and therefore operate at the full potential difference that exists across the load. Thus if the filaments of $V1$ and $V4$ were supplied by a single filament transformer winding, the common connection would short-circuit the load. Therefore, the filament windings of $V1$ and $V4$ must be insulated from each other to withstand the full output voltage across the load. If the lower end of the load, R, is grounded, filament windings for $V1$ and $V4$ operate alternately at the full difference of potential of the high-voltage secondary with respect to ground. Thus $V1$ and $V4$ must be well insulated from ground also.

This disadvantage does not apply to the rectifier elements of the dry-disk type which is often used in bridge circuits.

Filter Circuits

The preceding paragraphs have discussed means of converting alternating current into pulsating direct current. Most electronic equipments require a smooth d-c supply,

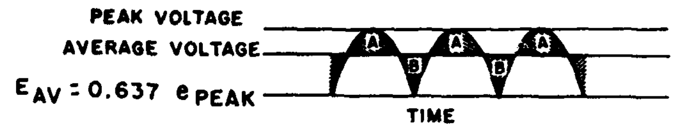


Figure 3–10.—Unfiltered output voltage of a full-wave rectifier.

approaching the ripple-free output of a battery. Conversion of pulsating direct current to pure direct current is accomplished by the use of properly designed filters.

The unfiltered output of a full-wave rectifier is shown in figure 3–10. The polarity of the output voltage does not

reverse, but its magnitude fluctuates about an average value as the successive pulses of energy are delivered to the load. In figure 3–10, the average voltage is shown as the line that divides the waveform so that area A equals area B. The fluctuation of voltage above and below this average value is called RIPPLE. The frequency of the main component of the ripple for the full-wave rectifier shown in figure 3–10 is twice the frequency of the voltage that is being rectified. In the case of the half-wave rectifier the ripple has the same frequency as the input alternating voltage. Thus, if the input voltage is obtained from a 60-cps source, the main component of ripple in the output of a half-wave rectifier is 60 cycles per second, and in the full-wave rectifier it is 120 cycles per second.

The output of any rectifier is composed of a direct voltage and an alternating or ripple voltage. For most applications, the ripple voltage must be reduced to a very low amplitude. The amount of ripple that can be tolerated varies with different applications of electron tubes.

The PERCENTAGE OF RIPPLE is 100 times the ratio of the rms value of the ripple voltage at the output of a rectifier or filter to the average value, E_o, of the total output voltage. Figure 3–11 indicates graphically how the percentage of

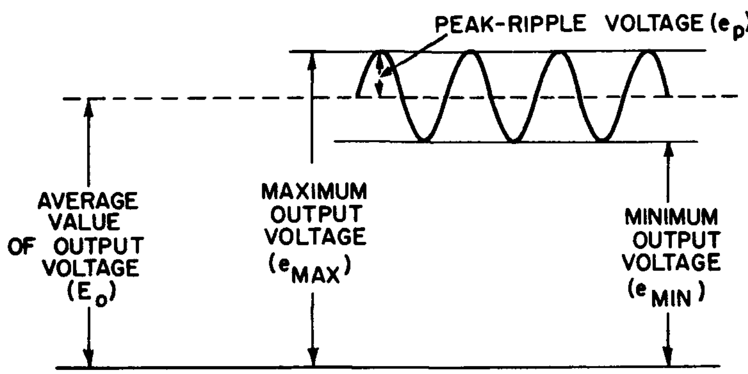


Figure 3–11.—Percentage of ripple.

ripple may be determined. It is assumed that the ripple voltage is of sine waveform. The formula for determining the percentage of ripple is

$$\text{percentage of ripple} = \frac{E_{rms}}{E_o} \times 100,$$

where $E_{rms} = 0.707$ of e_p, and e_p is the peak value of the ripple voltage.

A circuit that eliminates the ripple voltage from the rectifier output is called a FILTER. Filter systems in general are composed of a combination of capacitors, inductors, and in some cases resistors.

CAPACITANCE FILTER.—Ripple voltage exists because energy is supplied in pulses to the load by the rectifier. The fluctuations can be reduced considerably if some energy can be stored in a capacitor while the rectifier is delivering its pulse and allowed to discharge from the capacitor between pulses.

Figure 3–12, A, shows the output of a half-wave rectifier. This pulsating voltage is applied across a filter capacitor (C in fig. 3–12, B) to supply the load, R. Because the rate of charge of C is limited only by the reactance of the transformer secondary and the plate resistance of the tube in the rectifier, the voltage across the capacitor can rise nearly as fast as the half-sine-wave voltage output from the rectifier. In other words, the RC charge time is relatively short. The capacitor, C, therefore, is charged to the peak voltage of the rectifier within a few cycles. The charge on the capacitor represents a storage of energy. When the rectifier output drops to zero, the voltage across the capacitor does not fall immediately. Instead, the energy stored in the capacitor is discharged through the load during the time that the rectifier is not supplying energy (when the anode is negative). The voltage across the capacitor (and the load) falls off very slowly if it is assumed that a large capacitance and a relatively large value of load resistance are employed. In other

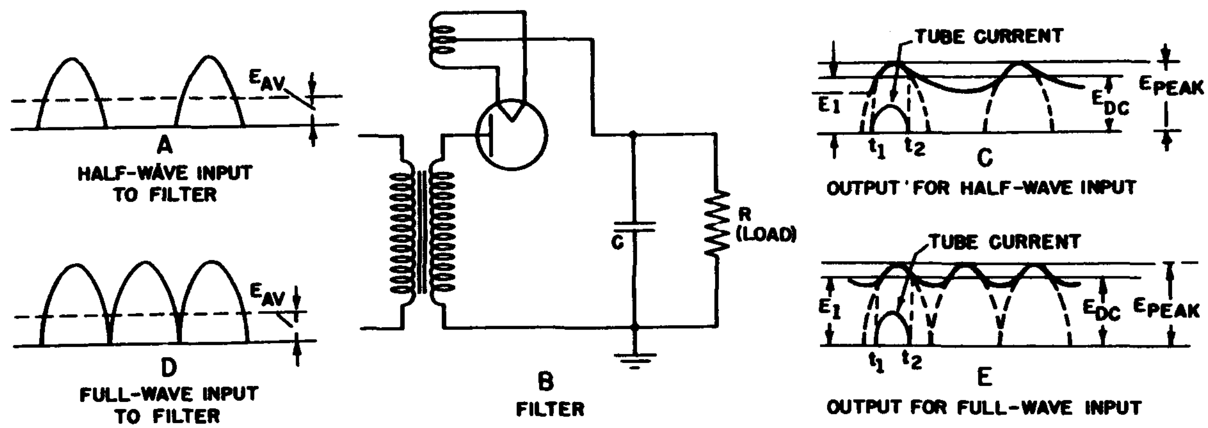


Figure 3-12.—Capacitance-type filter and waveforms.

words, the RC discharge time is relatively long. The amplitude of the ripple therefore is greatly decreased, as may be seen in figure 3–12, C.

Figure 3–12, D, shows the input voltage to the filter when a full-wave rectifier is used, and figure 3–12, E, shows the resultant output-voltage waveform.

After the capacitor has been charged (with either half-wave or full-wave input), the rectifier does not begin to pass current until the output voltage of the rectifier exceeds the voltage across the capacitor. Thus, in figure 3–12, C and E, current begins to flow in the rectifier when the rectifier output reaches a voltage equal to the capacitor voltage. This occurs at some time, t_1, when the rectifier output voltage has a magnitude E_1. Current continues to flow in the rectifier until slightly after the peak of the half-sine wave, at time t_2. At this time the sine-wave voltage is falling faster than the capacitor can discharge. A short pulse of current, beginning at t_1 and ending at t_2, is therefore supplied to the capacitor by the power source.

The average voltage of the rectifier output is shown in figure 3–12, A and D. Because the capacitor absorbs energy during the pulse and delivers this energy to the load between pulses, the output voltage can never fall to zero. Hence, the average voltage of the filtered output (fig. 3–12, C and E) is greater than that of the unfiltered input (fig. 3–12, A and D). However, if the resistance of the load is small, a heavy current is drawn by the load and the average or direct voltage falls. For this reason, the simple capacitor filter is not used with rectifiers that must supply a large load current. Also the input capacitor acts like a short circuit across the rectifier while the capacitor is charging. Because of this high peaked load on the rectifier tubes, the capacitor input filter is seldom used with gas tubes in high-current installations.

INDUCTANCE FILTER.—Because an inductor resists changes in the magnitude of the current flowing through it, an inductor can be placed in series with the rectifier output to help prevent abrupt changes in the magnitude of the current.

An inductance-type filter together with its input and output waveforms is shown in figure 3–13. The input waveforms from a half-wave and a full-wave rectifier are shown respectively in figure 3–13, A and B. Figure 3–13, C, shows the inductance-type filter, and figure 3–13, D and E, shows the output current for the half-wave and full-wave input respectively. When no inductor is used in series with R, the output current waveforms are indicated by dotted lines. The solid lines indicate the output-current waveforms when an inductor is used. The use of an inductor prevents the current from building up or dying down quickly. If the inductance is made large enough, the current becomes nearly constant.

The inductance prevents the current from ever reaching the peak value that it reached without the inductance. Consequently, the output voltage never reaches the peak value of the applied sine wave. Thus, a rectifier whose output is filtered by an inductor cannot produce as high a voltage as can one whose output is filtered by a capacitor. However, this disadvantage is partly compensated because the inductance filter permits a larger current drain without a serious change in output voltage.

PI-SECTION FILTER.—The ripple voltage present in a rectifier output cannot be eliminated adequately in many

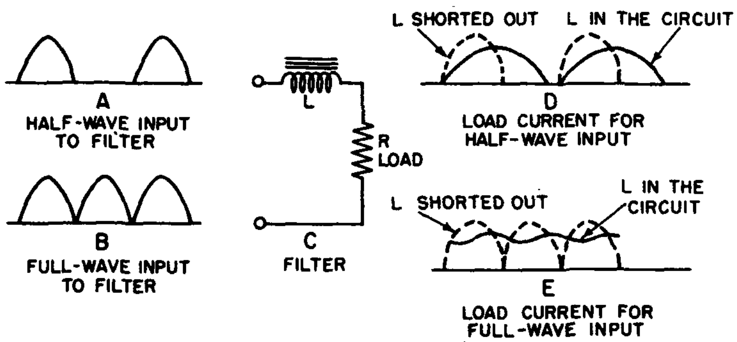


Figure 3–13.—Inductance-type filter and waveforms.

cases by either the simple capacitor or inductor filter. Filters that are much more effective can be made if both inductors and capacitors are used. The function of the capacitor is to store and release energy, while the inductors simultaneously tend to prevent change in the magnitude of the current. The result of these two actions is to remove the ripple from the rectifier output and to produce a voltage having a nearly constant magnitude.

Figure 3–14 shows a circuit diagram of an inductance-

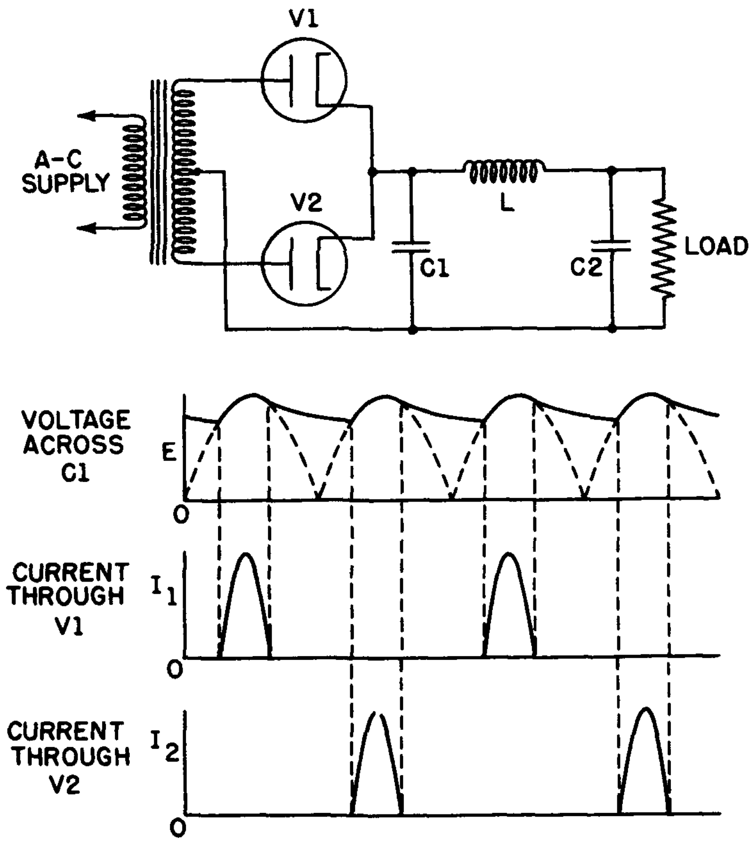


Figure 3–14.—Waveforms of current and voltage in rectifier with pi-section filter.

capacitance filter used primarily with receiver power supplies and other low-current power supplies.

This type of filter is given the name PI-SECTION because the configuration of the schematic diagram resembles the Greek letter π. It is also called a CAPACITOR INPUT FILTER. With this type of filter the output waveform closely approximates that of pure direct current. The first (input) capacitor acts to bypass the greatest portion of the ripple component to ground. In all filters the major portion of the filtering action is accomplished in this first component. The series choke in the pi-section filter serves to maintain the current at a nearly constant level during the charging and discharging cycles of the input capacitor.

At the bottom of figure 3-14 are shown the waveforms of current through $V1$ and $V2$ and the voltage across $C1$. The final capacitor, $C2$, acts to bypass residual fluctuations existing after filtering by the input capacitor and inductor. The current flow through the rectifier tubes is a series of sharp peaked pulses, because the input capacitor acts like a short circuit across the rectifier while the capacitor is charging. Because of this high peaked load on the rectifier tubes, the pi-section filter is used only in low-current installations such as radio receivers.

L-SECTION FILTER.—A second type of filter used primarily in high-current applications is the L-section filter, so named because of its resemblance to an inverted "L". A schematic diagram of this type of filter is shown in figure 3-15. The components perform the same functions as in the pi-section filter except that the inductor, or choke, input reduces the voltage output of the filter. This filter is also called a CHOKE INPUT FILTER. The input choke allows a continuous flow of current from the rectifier tubes rather than the pulsating current flow demanded by the capacitor input filter. The L-section filter is seldom used with half-wave rectifiers because there is no device to maintain current flow between half cycles.

Because of the uniform flow of current, the L-section filter

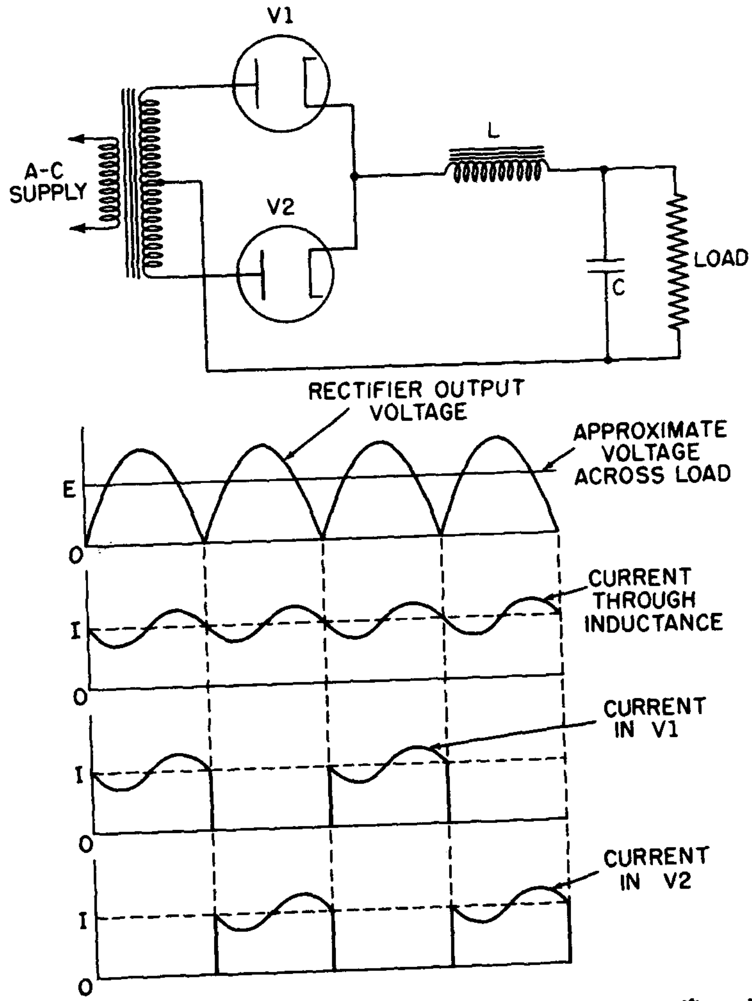


Figure 3-15.—Current and voltage waveforms in full-wave rectifier with L-section filter.

has applications in most high-power circuits, and is used with mercury-vapor rectifiers. It has the additional advantage of better voltage regulation. The inductive reactance of the choke reduces the ripple voltage without reducing the d-c output voltage.

Because the L-section filter is widely used in naval electronic equipment, the factors affecting the design of these filters are important to the electronics technician. As has been mentioned, this type of filter is generally used where high currents are required. In this respect, its advantage lies in the fact that it allows each rectifier tube to operate at a relatively constant level of current flow during its half-cycle of operation. This type of operation allows a rectifier to supply the maximum current to the load that it is capable of delivering. A disadvantage of the L-section filter is that instead of delivering a voltage equal to the peak value of the transformer secondary, it supplies a voltage equal to the AVERAGE of the a-c voltage delivered to the rectifier.

Two L-section filters are sometimes used in series to obtain a higher degree of filtering action.

VOLTAGE REGULATION.—The output voltage developed by any source of power tends to decrease when current is drawn from the source. The amount of change in the output voltage is usually expressed by a quantity called the PERCENTAGE OF VOLTAGE REGULATION.

The formula for the percentage of regulation is

$$\text{percentage of voltage regulation} = \frac{E_{NL} - E_{FL}}{E_{FL}} \times 100,$$

where E_{NL} is the no-load voltage, and E_{FL} is the output voltage when full-load current is flowing out of the supply. For example, assume the no-load voltage of a certain power supply to be 300 volts and the voltage at the output terminals to be 250 volts when the load resistance is applied and load current begins to flow. Substituting these values in this formula gives

$$\text{percentage of voltage regulation} = \frac{300-250}{250} \times 100 = 20 \text{ percent.}$$

The difference between the no-load voltage and the full-load voltage is caused by the flow of load current through the internal resistance of the power supply. The *IR* drop caused by the load current within the supply circuit is subtracted from the voltage available for the load resistance at the output terminals. A perfect power supply would have zero internal impedance and the percentage of regulation would be zero. Such a supply would provide the same voltage under full-load that it develops with no-load current flowing. In general, the lower the percentage of regulation, the better is the power supply in furnishing direct voltage and direct current for electronic equipment.

The regulation of the choke-input filter circuit is superior to that of the capacitor-input circuit as long as current is flowing in the filter choke. In this condition the output voltage changes very little when the load current changes in value. If, however, the load current should become zero, the choke coil can no longer prevent the first capacitor in the filter from charging to a value equal to the peak value of the applied voltage.

If the load current is a low value, or if it varies between a low value and zero, the regulation of the circuit is poorer than when larger currents are being drawn by the load. In order to improve the regulation of the choke-input filter, a resistor is often connected across the output terminals so that at least a minimum current will always flow through the choke.

Voltage Regulators

Most electronic gear used in the Navy can operate satisfactorily with a certain amount of variation in the supply voltage without suffering severe operational deficiency. However, some circuits are very critical and even a slight deviation from the normal supply voltage will cause unsatis-

factory operation. These circuits require the use of some type of voltage-regulating device.

A voltage-regulating device may be inserted in the circuit at one of two points—either between the rectifier and its load or at the power source that supplies electrical energy to the rectifier. The regulators that are used within a power supply are generally electronic and those affecting the power source itself (for example, the generator) are generally mechanical. Mechanical voltage regulators are treated in basic electricity texts.

FUNDAMENTAL VOLTAGE REGULATOR.—The regulator that is used to stabilize the output voltage of a rectifier usually takes the form of a variable resistance in series with the output. This variable resistance and the load resistance form a voltage divider. The variable element is controlled so that the voltage across the load is held constant.

Figure 3–16 shows a simple circuit that demonstrates this

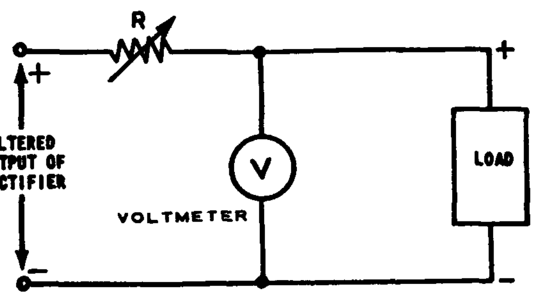


Figure 3–16.—Fundamental voltage regulator.

principle. The variable resistor, R, and the resistance of the load comprise a voltage divider that is connected across the rectifier output terminals. All the load current passes through R and causes a voltage drop across it. If the rectifier output voltage rises, the voltage across the load rises in proportion. To counteract this rise, the resistance of R is increased (manually) so that a greater proportion of the available voltage appears across R. The voltage across the

load therefore is held constant if the resistance of *R* is increased sufficiently to neutralize the increase of the rectifier output. If the resistance of the load increases, a greater fraction of the available voltage appears across the load. Therefore, the resistance of *R* must be increased in order to hold the voltage across the load constant.

In the system shown in figure 3–16, the resistor, *R*, must be varied manually in order to keep the voltmeter reading constant. If the voltmeter reading increases, *R* must be increased; if the voltmeter reading decreases, *R* must be decreased. This same type of action must take place in all of the voltage regulators that are to be discussed, but the action is automatic. The more complicated circuits that follow are more desirable than the simpler circuits because they are more accurate and can respond more quickly.

All the voltage regulators discussed in this chapter are essentially voltage dividers. The variable voltage drop may be supplied in many ways, but the action of most of the circuits may be explained in terms of the fundamental circuit shown in figure 3–16.

AMPERITE VOLTAGE REGULATOR.—A regulator tube that consists of an iron wire enclosed in a hydrogen-filled envelope is called an AMPERITE TUBE or BALLAST TUBE.

An amperite regulator circuit is shown in figure 3–17. The resistance of the iron wire in the ballast tube varies as the current through it changes. If the output voltage tends

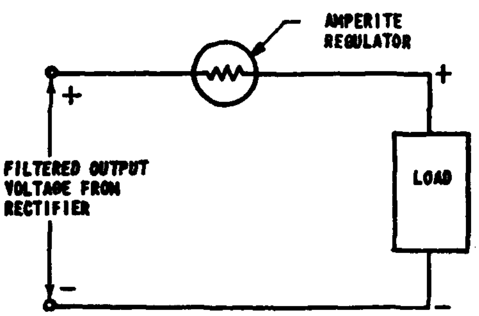


Figure 3–17.—Amperite regulator circuit.

to increase, more current flows through the ballast tube. The resistance of the tube then increases and more of the voltage drop takes place across the tube. Therefore, the voltage across the load remains nearly constant.

The amperite regulator does not regulate the voltage if the load changes. If the load increases, more current is drawn from the power supply and the load voltage falls. In addition, the greater current drawn causes the resistance of the amperite to increase, and the load voltage is made even lower by this additional drop.

Although the ballast tube may be used to compensate for line voltage variations it is generally inserted in series with several additional elements through which it is desired to maintain a constant current. In such applications, the resistance of the ballast tube changes to counteract the effect of changing voltage across the circuit.

GLOW-TUBE VOLTAGE REGULATOR.—In a glow-discharge tube, such as the neon glow tube, the voltage across the tube remains constant over a fairly wide range of current through the tube. This property exists because the degree of ionization of the gas in the tube varies with the amount of current that the tube conducts. When a large current is passed, the gas is very highly ionized and the internal impedance of the tube is low. When a small current is passed, the gas is lightly ionized and the internal impedance of the tube is high. Over the operating range of the tube, the product of the current through the tube and the internal impedance of the tube is practically constant.

A simple glow-tube regulator is shown in figure 3–18, A. The load current and the current that flows in the neon glow tube both pass through the series resistor, *R*. If the supply voltage drops, the voltage across the neon tube tends to drop. Therefore, the gas in the neon tube deionizes slightly and less current passes through the tube. The current through *R* is decreased by the amount of the current decrease in the tube. Because the current through *R* is less, the voltage drop across *R* is less. If the resistor is of the proper value relative to the load and to the glow tube that is used, the voltage across the

load is held fairly constant. In any case, the value of R must not be so large that the neon tube fails to ionize.

Glow tubes are designed to operate at various useful values of voltage. These values are usually indicated in the tube-type number.

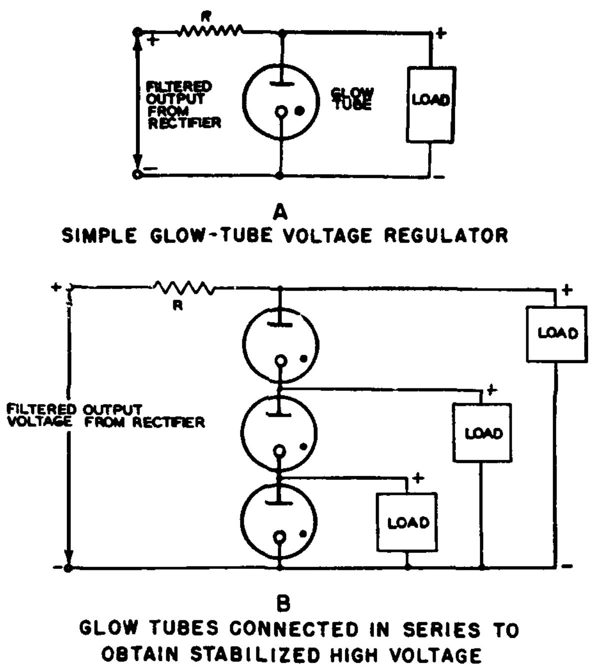


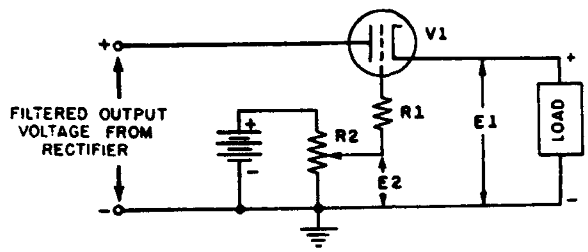
Figure 3–18.—Glow-tube regulators.

When a regulated voltage in excess of the maximum rating of one glow tube is required, two or more tubes may be connected in series, as in figure 3–18, B. This arrangement permits several regulated voltages with small current drains to be obtained from a single rectifier power supply.

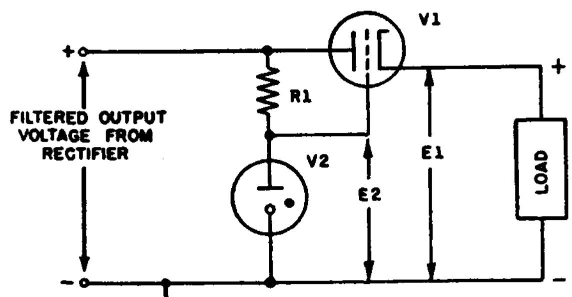
Electron-tube voltage regulator.—An electron tube may be considered as a variable resistance. When the tube is passing a direct current, this resistance is simply the plate-to-

cathode voltage divided by the current through the tube and is called the d-c plate resistance, r_p. For a given plate voltage, the value of r_p depends upon the tube current, and the tube current depends upon the grid bias

The variable resistor, *R*, of figure 3–16 can be replaced by an electron tube (fig. 3–19, A), because the electron tube



A
ELECTRON-TUBE VOLTAGE REGULATOR EMPLOYING A BATTERY FOR THE FIXED BIAS



B
ELECTRON-TUBE VOLTAGE REGULATOR EMPLOYING A GLOW TUBE FOR THE FIXED BIAS

Figure 3–19.—Electron-tube voltage regulator.

has a variable resistance. The effective resistance of *V*1 is established initially by the bias on the tube. Assume that the voltage across the load is at the desired value. Then the cathode is positive with respect to ground by some voltage,

E_1. The grid can be made positive relative to ground by a voltage, E_2, that is less than E_1. The potentiometer, $R2$, is adjusted until the bias (grid-to-cathode voltage), which is E_1-E_2, is sufficient to allow $V1$ to pass a current equal to the desired load current. With this bias, the resistance of $V1$ is established at the proper value to reduce the rectifier output voltage to the desired load voltage.

If the rectifier output voltage increases, the voltage at the cathode of $V1$ tends to increase. As E_1 increases, the negative bias on the tube increases and the effective plate resistance of the tube becomes greater. Consequently, the voltage drop across $V1$ increases with the rise in input voltage. If the circuit is properly designed, the increased voltage drop across $V1$ is approximately equal to the increase in voltage at the input to the regulator. Thus the load voltage remains essentially constant

The resistor, $R1$, is used to limit the grid current. This is necessary in this particular circuit because the battery is not disconnected when the power is turned off. However, the battery can be eliminated from the circuit by the use of a glow tube, $V2$, in figure 3–19, B, to supply a fixed bias for the grid of the tube. The action of the circuit is the same as the action of the circuit in figure 3–19, A.

The output voltage of the simple voltage regulators shown in figure 3–19 cannot remain absolutely constant. As the rectifier output voltage increases, the voltages on the cathode of $V1$ must rise slightly if the regulator is to function.

The voltage regulators shown in figure 3–19 compensate not only for changes in the output voltage from the rectifier, but also for changes in the load. For example, in figure 3–19, B, if the load resistance decreases, the load current will increase. The load voltage will tend to fall because of the increased drop across $V1$. The decrease in load voltage is accompanied by a decrease in bias voltage on $V1$. The bias voltage on $V1$ is equal to E_1-E_2. Thus, the effective resistance of $V1$ is reduced at the same time the load current is increased. The IR drop across $V1$ increases only a slight amount because R decreases about as much as I increases.

Therefore, the tendency for the load voltage to drop when the load is increased is checked by the decrease in effective resistance of the series triode.

Improved voltage regulator.—A very stable voltage regulator (more stable than those shown in fig. 3–19) can be designed by taking advantage of the high amplification possible with a pentode. A voltage regulator employing this type of tube is shown in figure 3–20. It produces an output that is independent of fluctuations in the a-c supply and changes in load over a wide range.

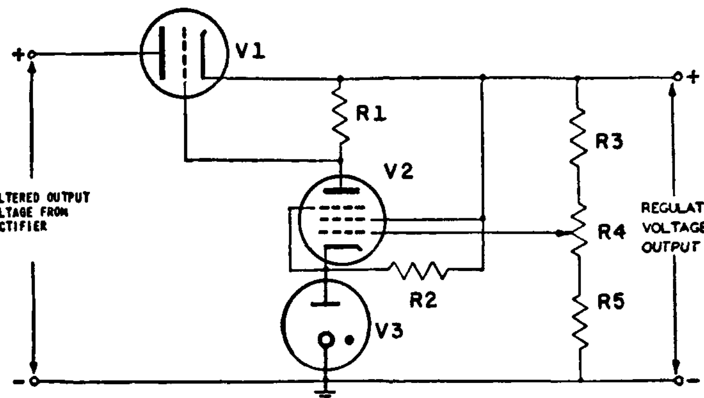


Figure 3–20.—Improved voltage regulator.

The output voltage of this regulator is developed across bleeder resistors $R3$, $R4$, and $R5$ in parallel with the resistance of the load. These resistors make up the resistance of one part of the total voltage divider. The other resistance, through which all of the load current must flow, is the cathode-to-plate effective resistance of $V1$. The other elements of the circuit are used to control the resistance of $V1$ and thereby to maintain a constant voltage across the load.

The potential of the cathode of $V2$ is held at a constant

positive value with respect to ground by the glow tube, $V3$. In other words, current flowing from ground through $V3$ causes an IR drop across $V3$ that maintains the cathode of $V2$ positive with respect to ground. The grid potential of $V2$ is a voltage selected by potentiometer $R4$. This potentiometer is set so that the grid-to-ground voltage is less positive than the cathode-to-ground voltage by an amount (the bias) that causes $V2$ to pass a certain plate current. In other words, the IR drop between the moving contact on $R4$ and ground is less than the IR drop across $V3$ by an amount that is equal to the bias on $V2$. The plate current of $V2$ flows through $R1$ and causes a drop across it. The magnitude of the voltage across $R1$ is the bias on tube $V1$. Therefore, the adjustment of potentiometer $R4$ establishes the normal resistance of $V1$. This adjustment is used to set the value of load voltage that the regulator is to maintain.

If the load voltage tends to rise, whether from a decrease in load current or from an increase in the input voltage, the voltage between the moving contact on $R4$ and ground will increase. The difference in this voltage and the fixed voltage across $V3$ decreases. These two voltages are in opposition, and the voltage between the moving contact on $R4$ and ground is less than the fixed voltage across $V3$. Thus, the grid bias of $V2$ decreases, and the plate current of $V2$ increases through $R1$. The increase in voltage across $R1$ increases the effective resistance of $V1$. If the load voltage tends to rise because of an increase in input voltage, this increase is accompanied by an increase in voltage across $V1$ and the rise in load voltage is checked. If the rise in load voltage is caused by a decrease in load current, this rise is checked because the IR drop across $V1$ remains constant, because the decrease in I is accompanied by an equal increase in R.

A pentode is used for $V2$ because of the high amplification possible with this type of tube. The use of such a tube makes the output voltage much more constant because small variations of load voltage are amplified sufficiently to cause operation of the circuit.

The anode of the glow tube, $V3$, is connected to the cathode of $V2$ and to the plus terminal of the regulated voltage output through resistor $R2$. It is necessary to connect the glow tube to the B+ in this way in order to cause the gas in this tube to ionize when the power supply is first turned on.

All the load current must pass through $V1$. For this reason, this tube must be capable of passing a large current. In some regulators a single tube does not have sufficient capacity to pass the required current. In such cases, several identical tubes may be connected in parallel.

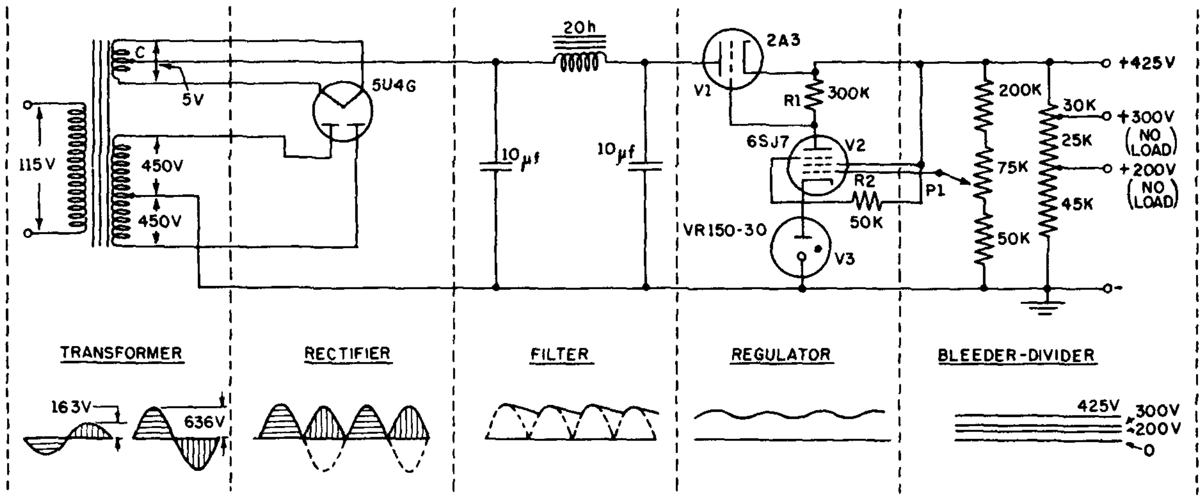
The type of regulator shown in figure 3–20 is used very widely to stabilize the output voltage of rectifier power supplies. Because of its excellent sensitivity to small changes in input voltage, this regulator is very effective in removing ripple from the output of rectifier power supplies. The regulator, then, serves also to filter the output of a rectifier, although the conventional filter systems usually are used in connection with a regulator.

Figure 3–21 is a complete rectifier-power-supply circuit, showing the power transformer, the full-wave rectifier tube, the filter circuit, the voltage regulator, and the voltage divider. This figure summarizes much of the foregoing power-supply discussion. (See pages 136 and 137.)

Voltage Dividers and Bleeders

A resistor is frequently placed across the output terminals of a rectifier power supply to bleed off the charge on the filter capacitor when the rectifier is turned off, or to apply a fixed load to the filter and thus improve the voltage regulation of the power supply. In the latter case, the resistor is designed to draw at least 10 percent of the full-load current in order for the change in power supply current to be less for a given change in load and thus reduce the magnitude of the variation in output voltage. In both cases the resistor is called a BLEEDER RESISTOR. If leads are connected to the resistor at various points to provide a variety of voltages that are less than the total output voltage, the resistor is called a VOLTAGE DIVIDER.

Explanations of action of rectifier, filter, voltage regulator, and output voltage divider, together with waveforms, are given on page facing this circuit diagram.



Low voltage is stepped up by the transformer from 115 volts to 900 volts. Center tap provides a dividing point so that 450 volts are applied to each section of the 5U4G rectifier. The ends of the transformer alternately become positive and negative.

Center tap C on heater winding is used to force plate current to divide equally in each filament lead. If there is no center tap, a voltage divider of two equal 50 ohm resistors may be put across the secondary to produce the same effect.

Alternately positive and negative voltage is applied to the plates of the rectifier.

The two plates conduct alternately as each plate is made positive in turn by the secondary of the transformer. Pulses of current flow from the filament line to each plate in turn. The plates alternately become positive and negative with the applied a. c., but the filament line will show a one-directional flow.

Capacitors charge when the rectifier conducts, and they discharge through the bleeder resistor when the tube is not conducting.

Choke builds up a magnetic field when the tube draws current. The field collapses as current decreases, tending to keep a constant current flowing in in the same direction through the bleeder resistor and the load.

Capacitor input (illustrated) gives higher voltage output with low current loads.

Choke input gives steadier output with less ripple under load conditions.

If the load draws more current or if the a-c input voltage falls, the terminal voltage of the power supply falls.

Resistor R1, tube V2, and gas-tube V3 are in series across the rectifier terminals. V3 holds the cathode of V2 at a constant positive potential with respect to ground, and setting of P1 determines bias on V2. A fall in terminal voltage causes more negative bias on V2, less current through V2, hence, less current through R1. Less IR drop across R1 causes less negative bias on V1. V1, then acts as a lower value resistor, and terminal voltage decrease is checked.

As a bleeder, the resistor is for safety to discharge the capacitors when power is removed.

As a load resistor, it acts as a stabilizer to protect the voltage regulator at no load, and to improve the regulation.

A voltage divider meets the requirements of a load resistor and a bleeder, but in addition has taps placed at intervals for voltage at less than the maximum.

It is usually grounded at the lower end but may be grounded at any higher point to get a negative output.

Figure 3-21.—Complete rectifier-filter-regulator-divider circuit.

Voltage divider circuits.—A resistor that is used as a load resistor may also serve as a divider and as a bleeder. A simple voltage divider composed of three similar resistors in series is shown in figure 3-22. As long as no load is drawn

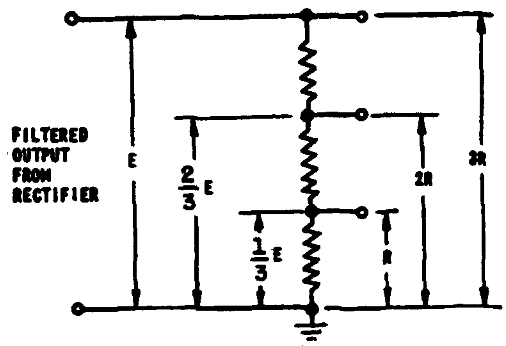


Figure 3-22.—Simple voltage divider.

from any terminal except the top, or line, terminal, the voltages across the resistors will divide in proportion to the resistance of each as shown in figure 3-22.

It is common practice to ground one side of voltage dividers. Therefore, ground potential is normally used as a reference for measurement of voltages as indicated at point *D* in figure 3-23, A. If a rectifier and its filter are connected so that no parts of the system are grounded, the divider can

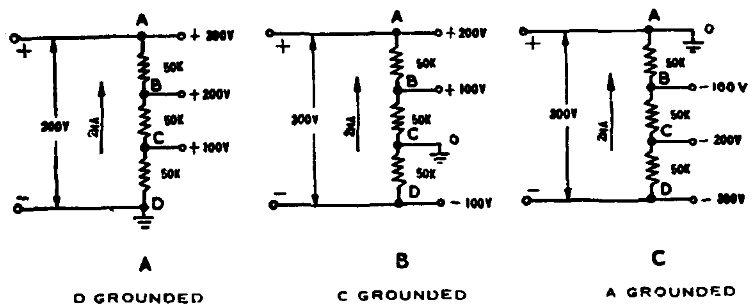


Figure 3-23.—Effect of moving the ground point on a voltage divider.

be grounded at any point without affecting the operation of the rectifier, provided the insulation of all parts is sufficient to withstand the voltage involved. Thus in figure 3–23, B, point C is grounded, and point D becomes negative with respect to ground. Such a circuit may be used to furnish both plate and bias voltages from the same power supply. In figure 3–23, C, point A is grounded, and all voltages along the divider are negative with respect to ground. Note, however, that point A will always be positive with respect to points, B, C, and D as long as the electron flow is maintained from D to A as shown in figure 3–23, C.

Operation of voltage dividers.—In the voltage divider circuits (fig. 3–23) it has been assumed that no load was attached to the divider except across terminals A and D, and that voltages could be measured without drawing appreciable current. As soon as a load is connected across the divider at any intermediate terminals, however, the voltage division shown in figure 3–23 is no longer correct. The resistance of the attached load forms a parallel circuit with that part of the divider across which it is placed, and therefore there is a change in the resistance of that part of the divider in relation to the total resistance between terminals A and D.

For example, in figure 3–24, a load of 150,000 ohms (150 k-ohms) is placed across BD, and a load of 50 k-ohms is placed across CD. The resistance between C and D is first determined by Ohm's law for parallel resistance—

$$R_{CD}=\frac{50\times 50}{50+50}=25 \text{ k-ohms.}$$

To this resistance is added the series resistance (50 k-ohms) between B and C, making a total of 75 k-ohms. The resistance across BD is then found by the parallel resistance rule. Thus,

$$R_{BD}=\frac{75\times 150}{75+150}=50 \text{ k-ohms.}$$

The total resistance between A and D before the main load is applied is the resistance between B and D plus the resistance between A and B, or $50+50=100$ k-ohms.

The total current, I, drawn through the divider and its

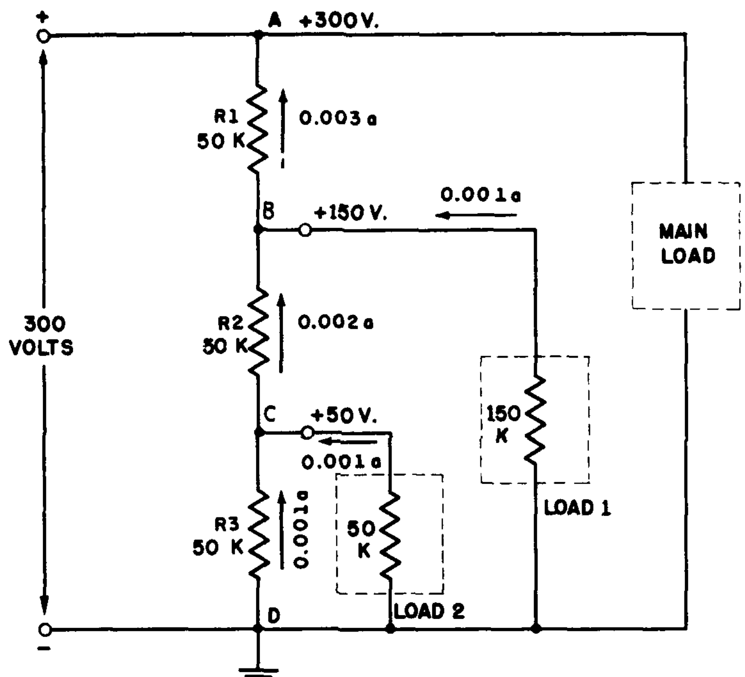


Figure 3-24.—Effects of loads on voltage division.

two loads is then the total voltage divided by the resistance between A and D. Therefore,

$$I=\frac{300}{100,000}=0.003 \text{ ampere.}$$

A current of 0.003 ampere flowing through $R1$ produces an IR drop of $50,000\times 0.003=150$ volts. Therefore, when load

1 and load 2 are connected as shown, the voltage across $R1$ increases from 100 to 150 volts. The voltage across load 1 is

$$300-150=150 \text{ volts.}$$

The current through load 1 is

$$\frac{150}{150{,}000}=0.001 \text{ ampere,}$$

and the current through $R2$ is

$$0.003-0.001=0.002 \text{ ampere.}$$

The 0.002 ampere flowing through $R2$ produces an IR drop of

$$50{,}000\times0.002=100 \text{ volts.}$$

Thus, the voltage remaining to be applied across CD is

$$150-100=50 \text{ volts.}$$

The current in load 2 is

$$\frac{50}{50{,}000}=0.001 \text{ ampere,}$$

leaving 0.001 ampere to flow through $R3$. As a check, the IR drop across $R3$ can be found as

$$50{,}000\times0.001=50 \text{ volts.}$$

Because this voltage is the same as that previously determined across CD, the value of current is correct.

Instead of a voltage of 200 volts between point B and ground, and 100 volts between point C and ground, as in figure 3–23, A, the voltage now is 150 volts at B and 50 volts at C, when the load values are as indicated in figure 3–24. Other load values will give correspondingly different values of voltage at B and C. Thus, the voltage appearing across

the intermediate terminals of a voltage divider divides proportionately to the values of the divider resistors only as long as no appreciable load current is drawn from these terminals. Under loaded conditions the voltages at these terminals will have various values, depending upon the resistance of the loads. A voltage divider must therefore be designed for the particular load conditions under which it is to operate.

VOLTAGE-MULTIPLYING CIRCUITS

Half-Wave Doubler

If a rectifier unit (a metallic type in this case) and a capacitor are connected in series across a source of alternating current, as shown in figure 3–25, the voltage will be doubled. The arrow in the doubler circuits in this chapter acts as the cathode, and electron flow is in the direction of this arrow. When the input voltage is as indicated in figure 3–25, A, the capacitor charges to the peak value of the line voltage. On the next half cycle the condition shown in figure 3–25, B, results. The full peak voltage across the capacitor is retained, but because of the polarity reversal of the source, the rectifier no longer conducts and as a result the voltage across the capacitor adds to that of the source. Therefore, the total voltage across the rectifier has a peak value twice that of the source. Thus, the output voltage varies between zero and twice the peak input voltage during each cycle.

The output voltage can be maintained over the entire cycle if a second rectifier unit and capacitor are added, as shown in figure 3–26. The second capacitor charges to twice the peak input voltage when rectifier *D*2 conducts and holds its charge during the time *D*2 is nonconducting.

Capacitor *C*2 cannot, however, maintain the full output voltage over the complete cycle if there is any appreciable load. This limitation results from the fact that when rectifier *D*2 is nonconducting, no current is drawn from the

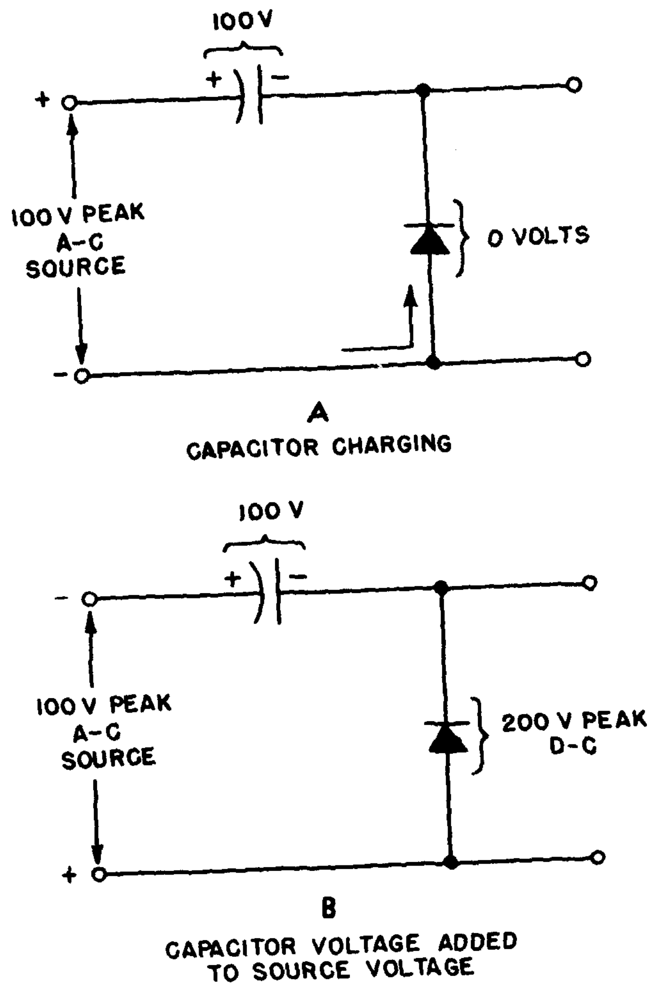


Figure 3-25.—Analysis of voltage doubler.

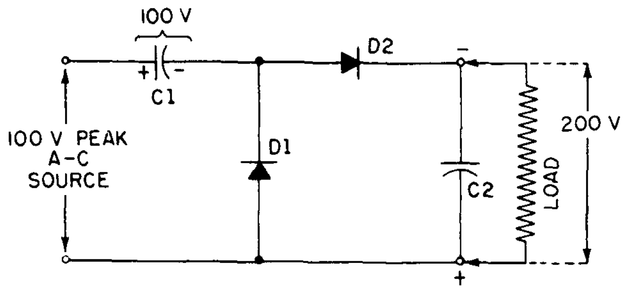


Figure 3–26.—Half-wave voltage doubler.

input circuit. Thus, *C*2 supplies the load current during discharge, and the output voltage falls proportionately. Because current is drawn from the source for only one-half of a cycle, this circuit is called a HALF-WAVE VOLTAGE DOUBLER. The half-wave characteristic and the size of capacitor *C*2 limit the use of this circuit to applications requiring only a small output current.

Full-Wave Doubler

A doubler that operates as a full-wave rectifier is shown in figure 3–27. Actually, such a connection is equivalent

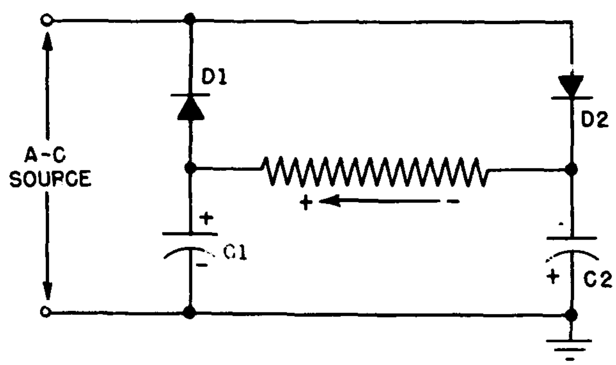


Figure 3–27.—Full-wave voltage doubler.

to connecting a pair of half-wave doublers across the source so that the direction of their conducting paths is opposite. As a full-wave rectifier, this circuit draws current from the voltage source during both halves of the input cycle. For one half cycle, *C*1 charges to the source voltage through rectifier *D*1, and for the next half cycle *C*2 charges to the source voltage through rectifier *D*2. The voltages impressed across *C*1 and *C*2 will therefore combine in series across the load to give the polarity and current flow indicated in the figure.

Voltage Multipliers

The process of increasing the voltage can be performed at higher levels of multiplication such as tripling and quadrupling. Theoretically, the voltage could be multiplied an infinite number of times by this process. Practical considerations, however, generally limit the multiplication to four or five times. Figure 3–28 shows a schematic diagram

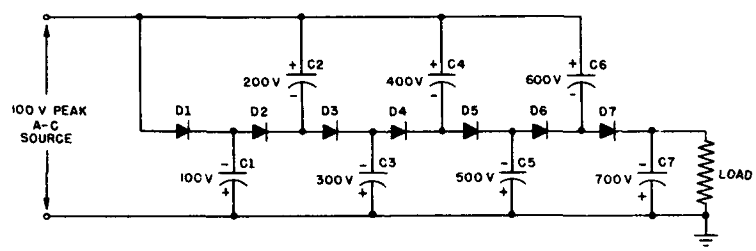


Figure 3–28.—Voltage multiplier.

in which an output voltage equal to seven times the peak of the input voltage is developed. The number of sections in this type of circuit could be extended to give high output voltages, but with each additional section the voltage regulation is adversely affected. The circuit acts as a half-wave multiplier and thus large capacitors must be employed to maintain current flow during alternate half cycles.

The operation of this circuit may be explained as follows: When the upper terminal is negative, electrons flow through all of the diodes, charging $C1$, $C3$, $C5$, and $C7$ to the peak voltage of the source. When the upper terminal is positive, the charges stored on these capacitors act in series with the input voltage to charge $C2$, $C4$, and $C6$ to twice the value of the input voltage. This reasoning may be continued through the first seven half cycles, and at this time the voltage across $C7$ will have been built up to seven times the input voltage.

Contrary to what may be expected upon first inspection of this circuit, the inverse peak voltage across any one of the rectifiers does not increase with the number of stages. Actually the peak inverse voltage across each rectifier, regardless of its position in the circuit, is twice the peak value of the input voltage.

Voltage multipliers are not widely used in naval electronic equipment. When high voltages are required, designers prefer the more conservative and dependable transformer and high-vacuum rectifier type of supply. When multipliers are utilized, the disk-type rectifier is well suited to the application because it requires no filament supply. If electron-tube rectifiers were used, the simplicity of the multiplier would be defeated by the necessity for providing a separate filament transformer for each stage of multiplication.

GRID-BIAS VOLTAGES

Grid Bias from B-Supply

In modern electronic equipment the grid-bias voltage is frequently derived from the plate supply voltage. Three methods are shown in figure 3–29.

The methods shown in figure 3–29, A and B, are commonly used in low-power amplifiers. The prime consideration is that the bypass capacitor be large enough to maintain a steady voltage across the resistor at the lowest operating frequency of the amplifier. In other words, the impedance

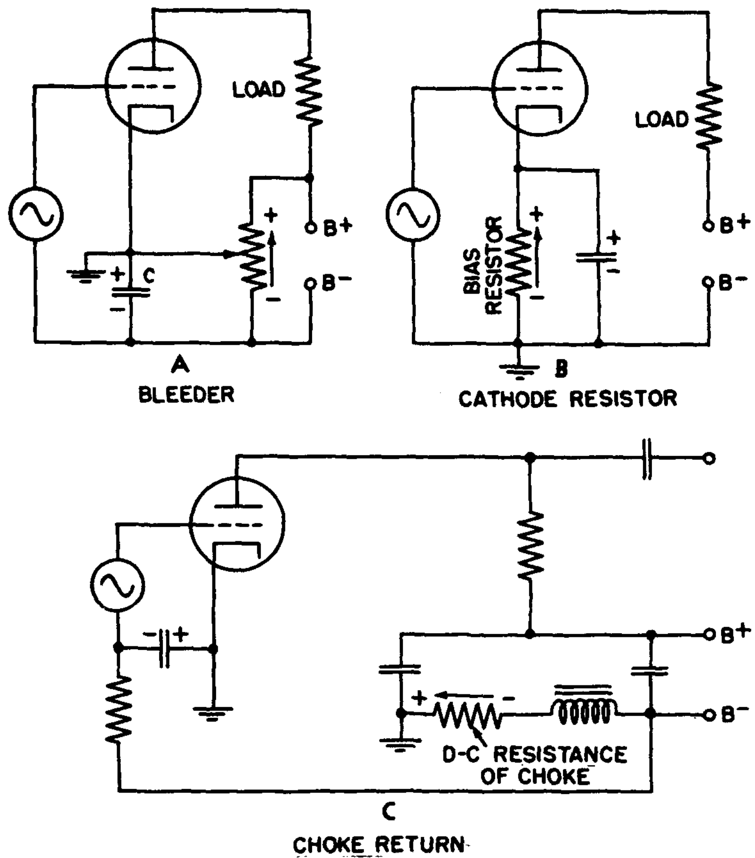


Figure 3-29.—Methods of obtaining grid-bias voltage from the plate power supply.

of the capacitor should be very low, compared with the resistance of the bias resistor, at the lowest frequency to be passed by the amplifier. These arrangements will obviously reduce the effective plate-to-cathode voltage by the absolute value of the bias voltage.

A brief consideration of current flow through the bleeder

and through the tube in figure 3–29, A, may be of value in understanding how the bias is established. Bleeder current flows from $B-$ to $B+$, and the no-signal (bias) voltage between grid and cathode is determined by the position of the movable contact. The grid is thus negative with respect to the cathode. The d-c component of plate current also flows through the lower portion of the bleeder. Capacitor C offers less opposition to the signal component of the plate current than does the lower portion of the bleeder. Therefore, the signal component flows from $B-$ through C to the cathode, and back to $B+$. Because the signal component does not flow through the bleeder the bias voltage is maintained at a steady value.

In figure 3–29, B, the bias is developed by the flow of the d-c component of the plate current through the bias resistor. The a-c component is passed around the bias resistor by the bypass capacitor. Thus, a steady negative bias is established between grid and cathode.

The method shown in figure 3–29, C, is used frequently in power amplifiers. This arrangement makes use of the resistance in the power-supply filter choke to supply the voltage drop necessary for grid bias. In this circuit the grid is connected to $B-$, and the signal component of the plate current flows from $B-$, through the low reactance of C, to the cathode, and back to $B+$ via the tube and load resistor.

Batteries are sometimes used for bias supplies when absolute stability of the bias voltage is necessary. Under these conditions the battery supplies no grid current and its effective life is its "shelf life." A more popular type of bias battery now in use is composed of tiny individual dry cells capable of being clamped together in a special type of holder resembling a fuse clamp. Any number of these dry cells can be placed in series to obtain bias voltages in multiples of 1.5 volts. These cells have extremely long shelf life.

Very large, high-power amplifiers frequently have separate bias supplies. They may be d-c generators or rectifier-filter systems. If d-c generators are used a filter must be placed in the output to eliminate any commutator ripple that may

be present. If a filter were not used, the ripple, no matter how small, would be amplified by the tube and cause distortion of the output signal.

Fixed Bias Voltage Supply

Some electronic equipments require a fixed bias. Transmitters, especially those that handle considerable amounts of power, need a fixed negative bias in order to protect the tubes and circuits, should other systems of bias fail. In the absence of a fixed bias a failure of the normal bias would cause a large increase in current through the tube and the associated equipment, and damage to the equipment might result.

One of the circuits that may be used to produce a fixed bias voltage is shown in figure 3-30. A conventional full-wave rectifier together with its filter circuit is shown at the top of the figure. This rectifier furnishes the high d-c voltages for

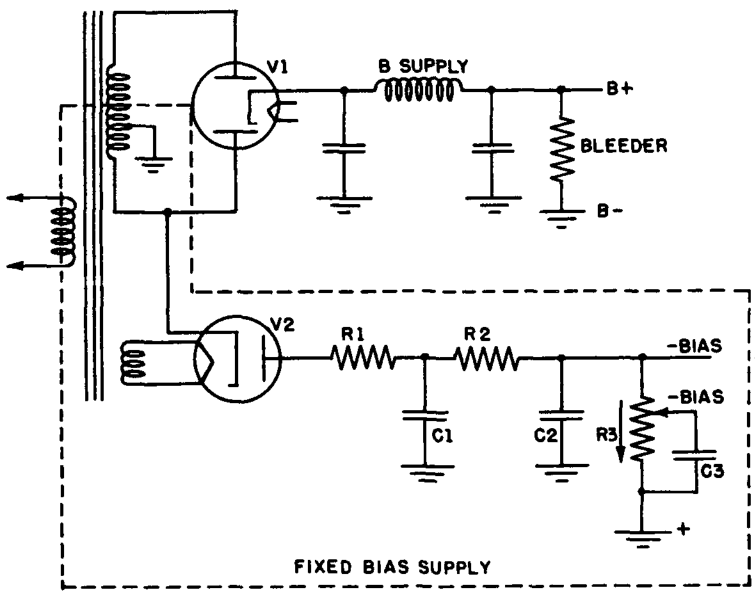


Figure 3-30.—Fixed bias supply circuit.

the plates and screen grids of the tubes. The fixed bias supply is shown enclosed in the dotted lines. The lower half of the high-voltage secondary is used, together with a half-wave rectifier diode, $V2$, and filters $R1$, $R2$, $C1$, and $C2$, to produce the fixed bias across $R3$.

When the lower end of the high-voltage secondary (fig. 3–30) is negative, electrons flow from the cathode to the plate of $V2$ and down through $R3$ to ground and thus back to the center tap on the secondary; thus the output voltage is negative with respect to ground in contrast with the B-supply circuit.

The tap on $R3$ permits the bias to be adjusted to the correct value. Bias resistor $R3$ generally has a relatively low value of resistance to minimize voltage variations due to grid current in class-C amplifiers, overdriven class-A amplifiers, or class-AB_2 amplifiers. If grid current should flow in one of the stages connected to the fixed bias supply, the voltage across the bias resistor (in the absence of capacitor $C3$) would increase and the bias would be increased accordingly.

When less bias voltage is needed than is developed in the circuit shown in figure 3–30, the 6-volt filament transformer supply circuit shown in figure 3–31, A, may be used instead.

The operation of the circuit in figure 3–31, A, may be explained by the use of the simplified circuit shown in figure 3–31, B. Assume that at a given instant point 1 is negative and point 2 is positive. Electrons flow from point 1 tnrough the tube and charge the right-hand plate of $C1$ negative. During this conducting time $R1$ is short-circuited by the tube. On the other half cycle when point 1 is positive and point 2 is negative, electrons cannot flow through the tube. Instead, $C1$, which has charged to approximately the full peak voltage of the supply, now partially discharges through $R1$, thus developing the bias voltage. The time constant of $C1R1$ is relatively long, so that only a small amount of charge leaks off $C1$ during the nonconducting period. $R2$ and $C2$ serve as filters, so that the voltage appearing between point 5 and ground is relatively free of ripple.

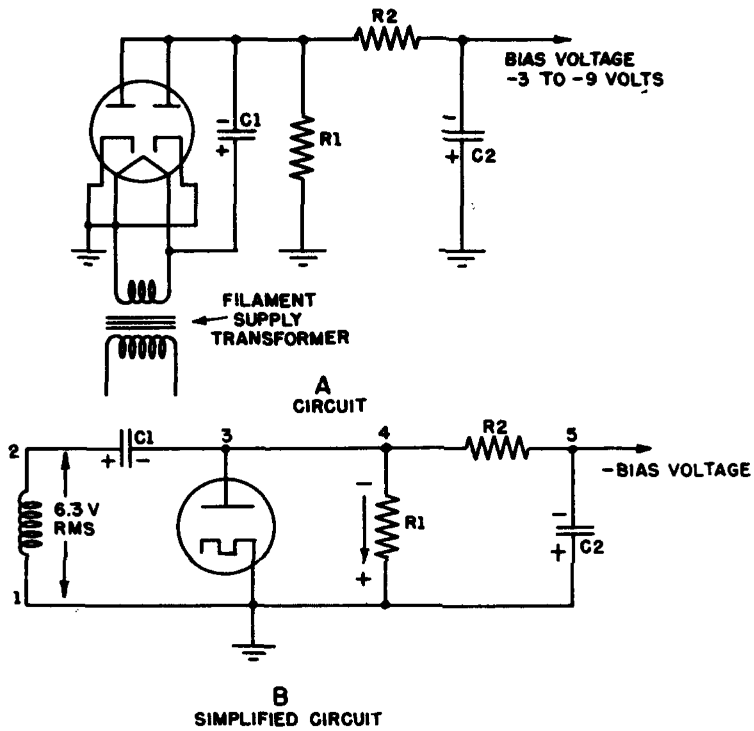


Figure 3-31.—Shunt-fed rectifier bias supply.

ELECTROMECHANICAL SYSTEMS

Dynamotors

The basic electrical power source in many aircraft is a 24-volt storage battery and an engine-driven generator. The generator charges the battery and supplies engine ignition, aircraft lights, and other electrical loads. In addition, aircraft communications equipment generally has incorporated within it another rotating machine called a DYNAMOTOR.

The dynamotor performs the dual functions of motor and generator, changing the relatively low voltage of the 24-volt

power supply into a much higher value for the plates and screens of electron tubes. The dynamotor usually employs two windings on a single armature. The two windings occupy the same set of slots and terminate in two or more separate commutators. The armature rotates in a single field frame with a conventional field winding to provide the exitation for both motor and generator. The motor armature winding is connected to the 24-volt power source and develops driving torque which rotates the armature as a motor. The generator winding is connected to the plates and screens of the electron tubes of the associated equipment and generates the relatively high voltage for these loads.

A functional diagram of a typical dynamotor is illustrated in figure 3–32. The heavy lines represent the motor circuit.

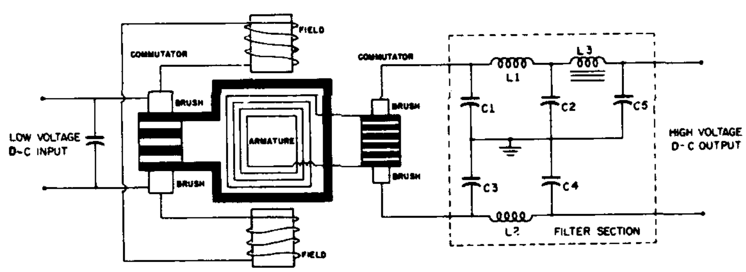


Figure 3–32.—Functional diagram of a dynamotor.

Relatively high current from the low-voltage source flows through the motor winding of the armature. The field is also energized from the low-voltage source. The interaction of the large current in the motor armature conductors with the field causes the armature to rotate.

The high-voltage winding, represented by the finer lines between the field, is wound in the same armature and rotates with the motor winding. When turning, the high-voltage winding cuts the lines of force of the common field and

generates a voltage which is developed across the brushes on the high-voltage commutator. The greater the number of turns in the high-voltage armature winding, the greater will be the voltage output.

Because the armature and field windings in the diagram of figure 3–32 are connected in parallel this is called a SHUNT-WOUND (shunt-connected) MOTOR. The desirable characteristic of this type of motor is that the speed remains fairly constant with changes in the load placed upon it.

The high current required by the motor necessitates a correspondingly large size in the motor components such as the commutator, brushes, and armature wire when compared to those components of the generator. The motor commutator is larger in diameter but has fewer segments than that of the generator. Because more turns in the generator armature winding produce a higher output voltage, there are a greater number of turns in that winding and the wire size is correspondingly reduced.

Filters are placed at the high-voltage output terminals to filter out high-frequency currents produced by sparking between the brushes and the rotating commutator segments, thereby eliminating any possible interference that the sparking may cause. The filter consists of a combination of chokes and capacitors such as shown in the typical filter section at the right of figure 3–32. The purpose of the chokes, *L*1 and *L*2, is to present a high impedance to the high-frequency currents so that they will not be present in the output. The low impedance of the capacitors, *C*1, *C*2, *C*3, and *C*4, bypasses high-frequency currents to ground.

Additional audio filtering is provided to eliminate commutator ripple which compares with the ripple found in the output of conventional a-c rectifiers. This audio filtering consists usually of a series inductor of comparatively high value and a shunt capacitor. It is represented in the figure by the iron-core choke, *L*3, and capacitor *C*5. The capacitor across the low-voltage input leads reduces sparking between the brushes and commutator at the input end of the dynamotor.

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Vibrators

The vibrator is another type of voltage supply used to obtain a high a-c or d-c voltage from a comparatively low d-c source. It has certain advantages over the dynamotor type of power supply. For example, the vibrator is lighter and less expensive than the dynamotor; it is also more efficient. However, the vibrator can be used only when a limited amount of high-voltage current is needed. Also, its life is relatively short, and it tends to produce radio interference (hash). It is extensively used in the "power packs" of lightweight mobile equipment. Actually, neither the dynamotor nor the vibrator are power supplies as such, being only the means by which low-voltage direct current is converted to high-voltage direct current.

A simple vibrator power supply is shown in figure 3-33.

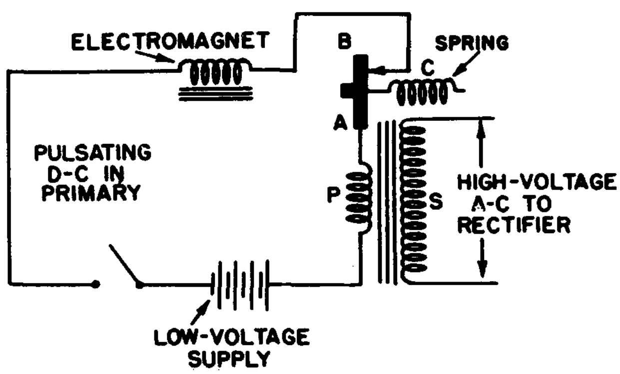


Figure 3-33.—Basic vibrator power-supply system.

It is nothing more than a simple interrupter, similar in many respects to a buzzer or doorbell. Pulsating direct current is used to energize the primary winding of a transformer which in turn induces an a-c voltage in the secondary. The turns ratio of the transformer windings are chosen to give the desired output voltage.

When the switch is closed in the circuit of figure 3–33, current flows from the battery through the electromagnet, and then through contact *B*, armature *A*, primary winding *P*, and back to the battery. In passing through the electromagnet the current sets up a magnetic field that attracts the armature. As the armature moves it breaks the circuit at contact *B*. As soon as the circuit is broken, the electromagnet no longer attracts the armature, thus allowing spring *C* to pull it back to the starting position. At the starting position, contact *B* again closes the circuit and the process is repeated. In this way a pulsating direct current that induces a high voltage in the secondary winding flows through the primary of the transformer.

The output voltage of the secondary is applied to a conventional rectifier and filter network, which converts the alternating current back into direct current, but at a higher voltage.

Two typical vibrator power-supply systems are shown in figure 3–34. In figure 3–34, A, is shown the nonsynchronous type of vibrator power supply. This type of power supply requires a separate rectifier and filter. Either cold-cathode or high-vacuum rectifiers are used with nonsynchronous vibrators.

The operation of this type of vibrator is much the same as that of the basic vibrator shown in figure 3–33. When the battery switch is closed, current flows through the lower half of the primary, the electromagnet, and back to the battery, producing an expanding magnetic field. As the armature is drawn down, the electromagnet is temporarily short-circuited, and loses its magnetism. The armature is released and makes contact with terminal 2. Current now flows through the upper half of the primary and back to the battery. At the same time this is occurring, the field previously established by the current in the lower half winding is collapsing. The effect of the simultaneous expansion of one field and collapsing of the other field is to increase the magnitude of the induced voltage in the secondary.

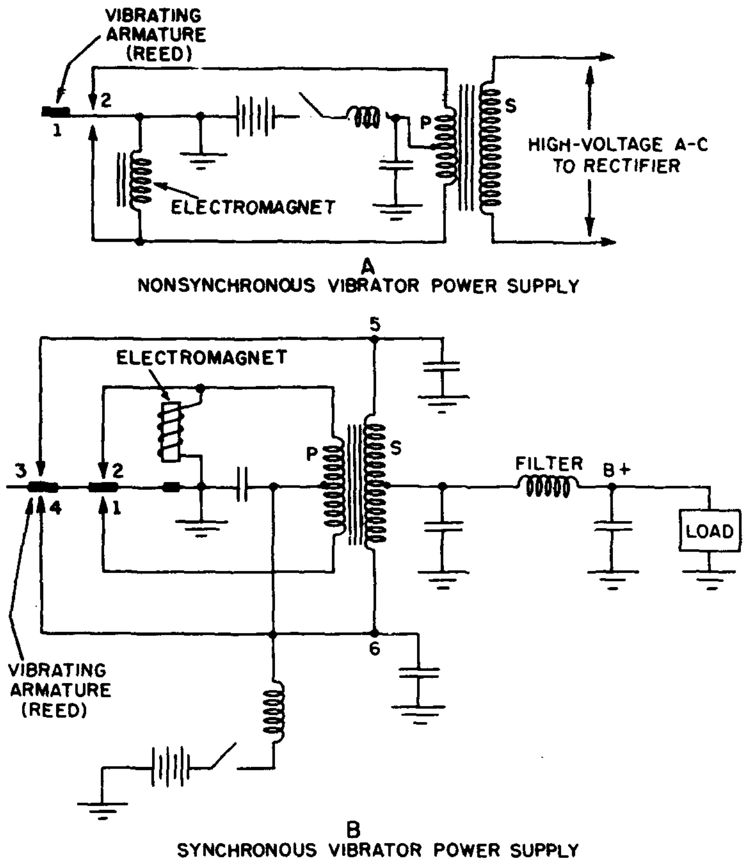


Figure 3-34.—Typical vibrator power-supply systems.

The synchronous vibrator, shown in figure 3-34, B, does not require a rectifier tube. This type of vibrator is called a synchronous vibrator because two additional contacts connected to the ends of the secondary winding are so synchronized with the contact in the primary circuit that rectification takes place.

Contacts 3 and 4 perform the function of rectification. Insofar as the primary is concerned, the action is similar

to that of the nonsynchronous vibrator. If at a given instant point 5 is negative and the armature is touching points 2 and 3, electrons will flow from point 5, to point 3, to ground; and return to the center tap on the secondary by way of the load. A half cycle later, point 6 is negative and the armature is touching points 4 and 1. Electrons then flow from point 6 through 4 to ground, and return to the center tap via the load. Thus, current always flows through the load in the same direction—that is, from ground to the center tap.

Inverters

In many naval aircraft the primary source of power that is available for use in the electronic equipment is the 24-volt d-c supply that is used to run dynamotors and to supply low-voltage filament and heater power of low-power transmitters and receivers. In addition, 115-volt alternating current is often derived from the low-voltage direct-current source by the use of a device called an INVERTER. The inverter is a rotating type of machine that takes an input of 24 volts d-c and changes it into 115 volts a-c at about 800 cycles per second.

The 115-volt power is then used to provide the input for rectifier power supplies in the high-power electronic equipment, such as radar units.

Electromechanical systems that change alternating current to direct current are commonly called CONVERTERS, and electromechanical systems that change direct current to alternating current are commonly called INVERTERS.

Because alternating current is used for some of the instruments on aircraft, and also for control equipment, radar, radio, fluorescent lighting, etc., some means of producing alternating current from the aircraft direct-current system is necessary. Although lighter equipment may be used in certain instances, the motor-generator (d-c motor, a-c generator) set is more flexible and in general more satisfactory from the point of view of frequency and voltage control.

Common frequencies of aircraft inverters are 400 and 800 cycles per second. The higher frequencies permit smaller inductive components in the equipment and thus the weight may be kept to a minimum.

Both d-c motors and a-c generators are treated in texts on basic electricity and hence will not be treated here.

QUIZ

1. How does the B-supply differ from the A-supply?
2. Why is it important in a directly heated cathode to return the grid and plate circuits of a tube to a point the exact electrical center of the filament circuit when a-c is used for heating?
3. Why is the center-tap grid and plate return not needed when indirectly heated cathodes are used?
4. What type of rectifier tube (high-vacuum or gas-filled) is most widely used in low-current applications?
5. What are two of the important characteristics of the high-vacuum rectifier tube?
6. Why is the mercury-vapor rectifier more efficient than the high-vacuum rectifier?
7. What is the normal voltage drop across a mercury rectifier tube when it is conducting?
8. Why is the peak inverse voltage rating of the individual copper-oxide rectifier units relatively low?
9. How may the peak inverse voltage rating of dry-disk rectifiers be increased?
10. Why is the cathode or figure 3–7 positive with respect to ground?
11. What are the rms and average values of the unfiltered voltage across a load supplied by a full-wave rectifier having a peak output voltage of 200 volts?
12. Why is the transformer inductance in a full-wave rectifier not reduced as it is in the half-wave rectifier?
13. With a given transformer, what is the relative output voltage of a bridge rectifier compared with that of a conventional full-wave rectifier?
14. Give one disadvantage of the bridge rectifier employing filament-cathode-type electron tubes?
15. Why is the charge time of C in figure 3–12 relatively short and the discharge time relatively long?
16. Why is a simple capacitor filter not used with rectifiers that must supply a large load current or with gas-tube rectifiers?
17. Why is the output voltage of a choke input filter lower than for an input capacitor filter having the same input voltage?
18. Where does the major portion of the filtering take place in a pi-type filter?
19. Why is the L-section filter seldom used with half-wave rectifiers?

20. What causes the difference between the no-load voltage and the full-load voltage of a power supply:
21. In a choke-input filter, why must a minimum amount of current be maintained at all times through the choke?
22. If in figure 3–16 the filtered output voltage from the rectifier should increase in value, would *R* have to be increased or decreased in value in order to maintain a constant voltage across the load?
23. How is the voltage regulation affected if an amperite voltage regulator is used and the load varies?
24. How does a glow-tube regulator (fig. 3–18) hold the load voltage constant when the source voltage changes?
25. When the negative bias on the grid of *V*1 in figure 3–19, A, increases what happens to the effective plate resistance of the tube?
26. In the electron-tube voltage regulator shown in figure 3–19, what checks the drop in load voltage when the load is increased?
27. What is the advantage of using a pentode in the improved voltage regulator in figure 3–20?
28. What are two functions of bleeder resistors?
29. What is the function of a voltage divider?
30. In a voltage divider what is the purpose of grounding some point other than an end terminal?
31. What is the result of applying a load across one portion of a voltage divider?
32. Why is the output current of the half-wave voltage doubler of figure 3–26 limited?
33. What is the purpose of connecting a large cathode bypass capacitor across the cathode-bias resistor in figure 3–29 when low frequencies are being amplified?
34. Why is a fixed bias used in transmitters that handle large amounts of power?
35. Why is the generator winding of a dynamotor that derives its power from a 24-volt system, composed of more turns than the motor winding?
36. What function is performed by a vibrator?
37. What distinguishes a synchronous vibrator from a nonsynchronous vibrator?
38. For what purpose are inverters used?

CHAPTER 4

INTRODUCTION TO ELECTRON-TUBE AMPLIFIERS

CLASSIFICATION OF AMPLIFIERS

The most important function of an electron tube is its ability to amplify or increase the amplitude of the input signal. An electron-tube amplifier consists of one or more tubes and associated circuit elements necessary for its operation and is used to increase the voltage, current, or power of a signal. For example, a minute amount of power at the input of a broadcast receiver is amplified by a number of amplifier stages in the receiver to the level necessary to operate a loudspeaker. The amount of amplification or gain that results is dependent primarily upon the number of amplifier stages used. GAIN is defined as the ratio of output to input. The greater the number of stages the greater the over-all gain will be. One stage may have a larger gain (gain per stage) than another. In many circuits the gain is dependent largely upon the amplification afforded by the electron tube. In other circuits the gain may be due to transformer action or the resonant qualities of a circuit. The amplification of the tube itself is expressed as an amplification factor, μ (defined in chapter 2).

A signal voltage, e_s, of sine waveform applied to the control grid of a tube (fig. 4–1, A) results in plate current variations through the load impedance and voltage variations

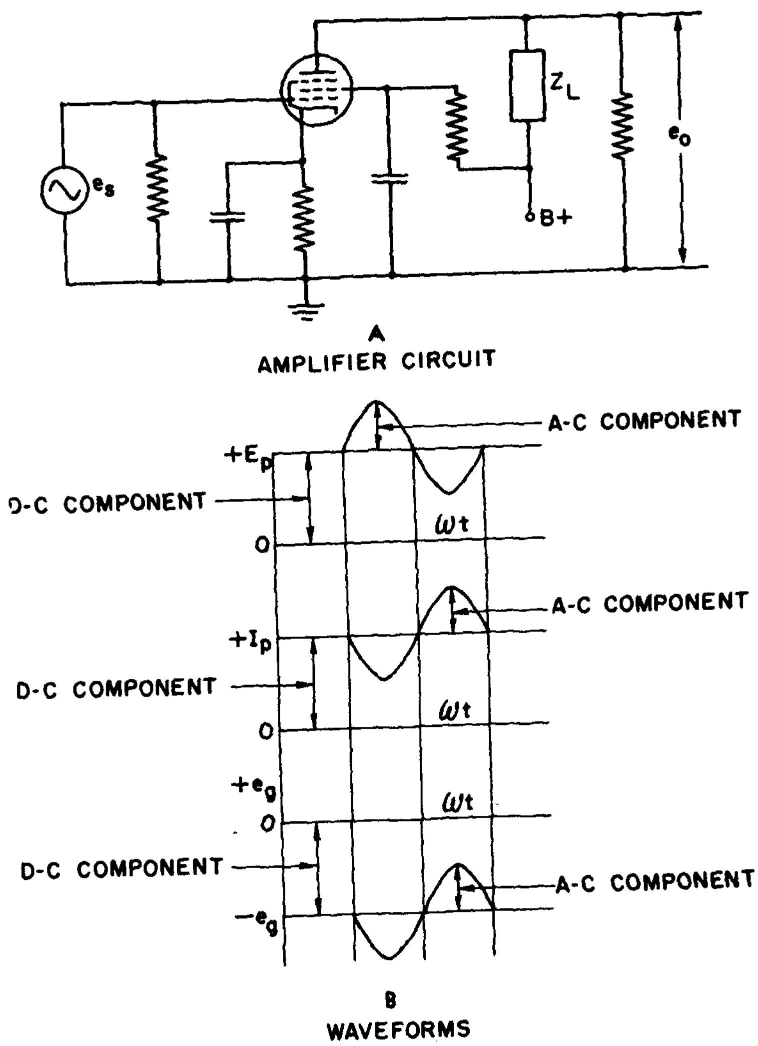


Figure 4-1.—One-stage amplifier and waveforms.

between plate and ground, as shown in figure 4–1, B. The voltages and currents are made up of a d-c component that exists when no signal is present and an a-c component that exists in addition to the d-c component when a signal is applied to the grid. In most cases the a-c component is of chief interest although the d-c component determines the portion of the tube characteristic in which the operation occurs. The a-c components of plate voltage and current constitute the useful output of the tube.

In practical applications, input and output coupling circuits must be used with the electron tube. If resistance-capacitance coupling is used, the voltage gain of the stage will be less than the amplification factor of the electron tube because of losses in the coupling elements. If transformer coupling is used, the voltage gain may be greater or less than μ depending on whether the coupling transformer has a step-up or step-down turns ratio.

In order to obtain certain waveform characteristics, amplifiers are sometimes purposely designed to distort the signal. When voltage or power is to be amplified without appreciably changing the shape of the wave, as in high-fidelity amplifiers, it is generally necessary to sacrifice some of the gain that the stage would normally have if this condition were not imposed.

Amplifiers may be classified in a number of ways such as according to use, bias, frequency response, or resonant quality of the load.

According to Use

When classified according to use or type of service, amplifiers fall into two general groups—VOLTAGE AMPLIFIERS and POWER AMPLIFIERS.

VOLTAGE AMPLIFIERS.—Voltage amplifiers are so designed that signals of relatively small amplitude applied between the grid and the cathode of the tube will produce large values of amplified signal voltage across the load in the plate circuit. In order to produce the largest possible amplified signal voltage across the plate load (which may be a resistor, an

inductor, or an impedor) this value of impedance must be as large as practicable.

The GAIN of a voltage amplifier is the ratio of the a-c output voltage to the a-c input voltage. This type of amplifier is commonly used in radio receivers to increase the r-f or i-f signal to the proper level to operate the detector. It is used also to amplify the a-f output of the detector stage. In phone transmitters, voltage amplifiers are used to increase the output of the microphone to the proper level to be applied to the modulator.

POWER AMPLIFIERS.—Power amplifiers are designed to deliver a large amount of power to the load in the plate circuit. Since power, in general, is equal to the voltages times the current, a power amplifier must develop across its load sufficient voltage to cause rated current to flow. The POWER AMPLIFICATION of such a circuit is the ratio of the output power to the input grid driving power.

The load impedance of a power amplifier is selected to give either maximum plate efficiency or maximum power output for a certain minimum level of distortion. PLATE EFFICIENCY is the ratio of useful output power (a-c voltage component times a-c current component times cos θ) to d-c input power to the plate (plate current times plate voltage).

POWER SENSITIVITY, another term used with power amplifiers, is the ratio of the power output in watts to the square of the effective value of grid signal voltage and is measured in mhos. Expressed as a formula

$$\text{power sensitivity} \quad \frac{\cos\theta}{e_g^{\,2}},$$

where E and I are the effective values of output voltage and current respectively, θ is the phase angle between them, and e_g is the effective value of the input signal voltage.

Power amplifiers are commonly used as the output stage of radio receivers. They are used also in transmitters to increase the power of the modulated carrier to the desired level before it is fed to the antenna.

According to Condition of Operation (Bias)

Amplifiers may be classified also according to the conditions under which the tube operates—that is, according to the portion of the a-c signal voltage cycle during which the plate current flows as controlled by the bias on the grid. The four classes of amplifier operation according to bias are class A, class B, class AB, and class C.

CLASS-A AMPLIFIERS.—Class-A amplifiers are biased so that, with normal input signal, plate current flows during

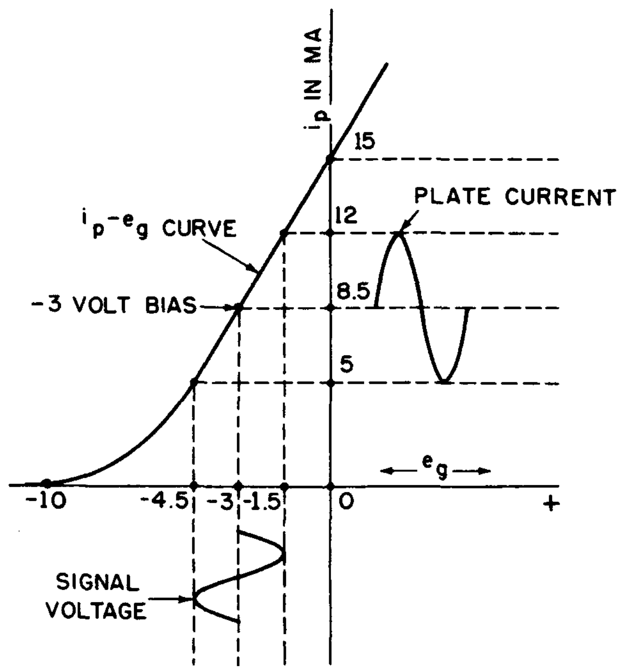


Figure 4-2.—Class-A operation.

the entire input cycle, and the amplification is essentially linear, as indicated in figure 4–2. Grid current does not flow in most class-A amplifiers.

To show that grid current does not flow during any part

of the input cycle, the subscript "1" may be added to the letter or letters of the class identification. The subscript "2" may be used to indicate that grid current flows during some parts of the input cycle. Thus if the grid is not driven positive at any time in the class-A cycle no grid current will flow and the amplifier is designated class A_1.

The principal characteristics of class-A amplifiers are minimum distortion, low power output for a given tube (relative to class-B and class-C amplifiers), high power amplification, and relatively low plate efficiency (20 to 35 percent). This type of amplifier finds wide use in various audio systems where low distortion is important.

CLASS-B AMPLIFIERS.—Class-B amplifiers are biased so that no plate current flows when no signal is applied to the grid. Plate current then flows for approximately one-half of each cycle of grid signal voltage (fig. 4–3). Such amplifiers are characterized by medium power output, medium plate efficiency (50 to 60 percent), and moderate power amplification. Since the a-c component of plate current is

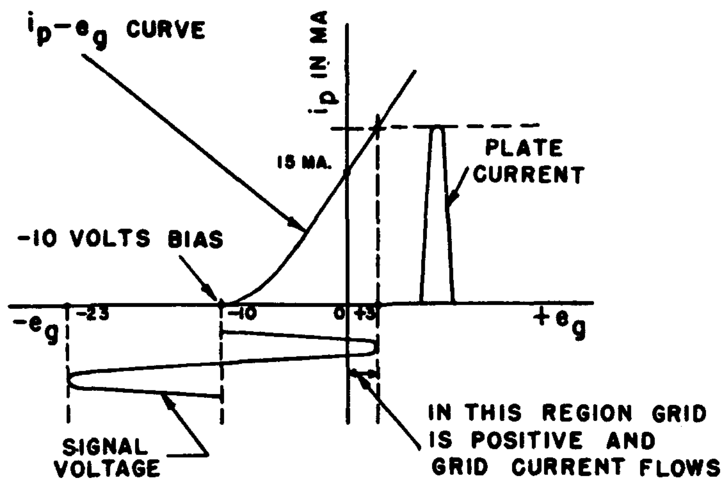


Figure 4–3.—Class-B operation.

proportional to the amplitude of the grid signal voltage, the output power is proportional to the square of this voltage.

Single-ended (single-tube) class-B amplifiers are used in r-f amplifier stages having a parallel-tuned circuit as the plate load. Two of these tubes may be used in push-pull output stages of audio-frequency amplifiers. In this circuit (fig. 4-4) each tube supplies that half of the waveform not supplied by the other. Thus, the resultant amplified wave-

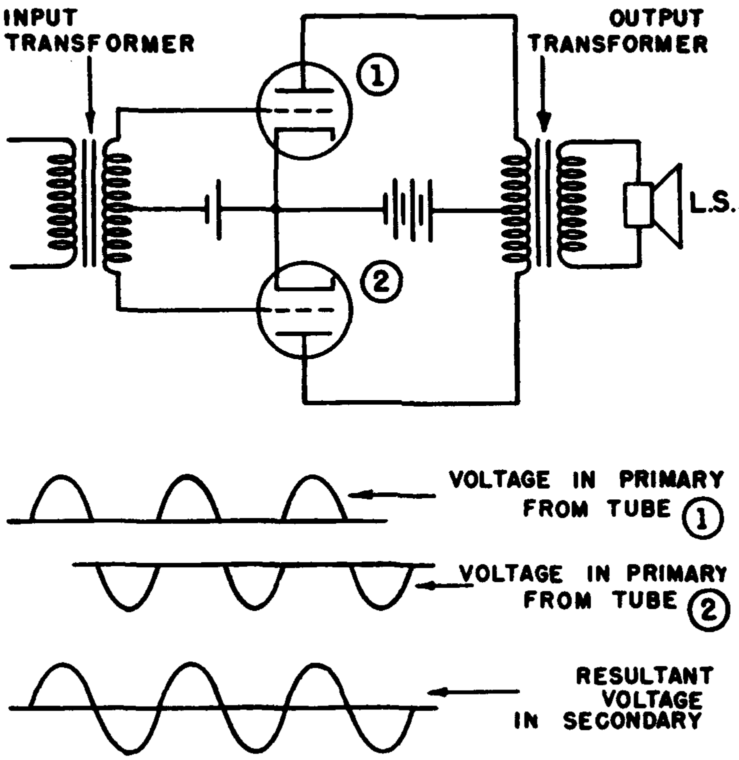


Figure 4-4.—Class-B push-pull amplifier.

form is a nearly true reproduction of the signal applied between the two grids. This circuit should not be confused with the full-wave rectifier circuits shown in the preceding

chapter since the load in the class-B amplifier is supplied with alternating current.

CLASS-AB AMPLIFIERS.—Class-AB amplifiers have grid biases and input-signal voltages of such values that plate current in a single tube flows for appreciably more than half the input cycle but for less than the entire cycle, as

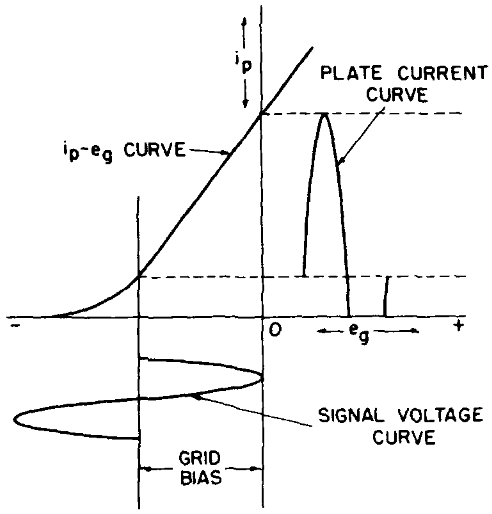


Figure 4–5.—Class-AB operation.

indicated in figure 4–5. Class-AB operation is essentially a compromise between the low distortion of the class-A amplifier and the high efficiency of the class-B amplifier.

If the input signal drives the grid positive with respect to the cathode, grid current will flow during the positive peaks and the amplifier is designated as a class-AB_2 amplifier. Although a class-AB_2 amplifier delivers slightly more power to its load, the class-AB_1 amplifier has the advantage of presenting to its driver a constant impedance. In contrast with this effect the amplifier that draws grid current

over a portion of its input cycle presents a changing impedance to its driver at the point where grid current starts to flow. Thus before grid current the impedance may be relatively high, and during that part of the input cycle when grid current flows the impedance falls to a relatively low value. The driver that supplies this kind of load must be designed to supply undistorted power to the load during these periodic intervals of low impedance.

CLASS-C AMPLIFIERS.—Class-C amplifiers have a bias that is appreciably greater than cutoff; consequently plate current in a single tube flows for appreciably less than half of each cycle of the applied grid signal voltage (fig. 4–6). This class

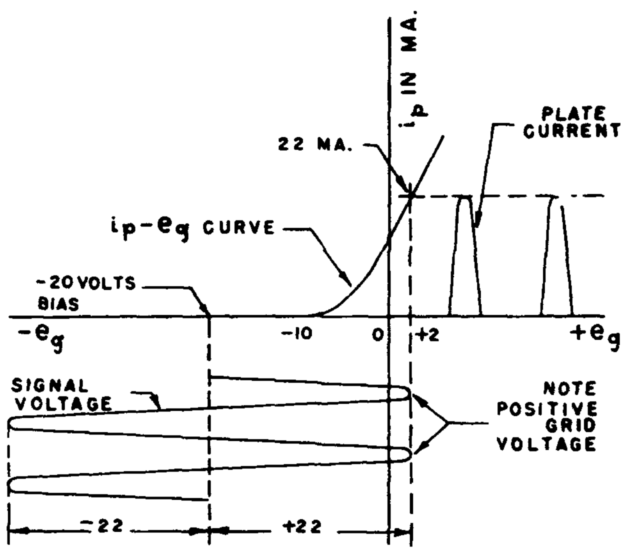


Figure 4–6.—Class-C operation.

of amplifier has a relatively high plate efficiency (70 to 75 percent), high power output, and low power amplification.

Class-C amplifiers are not used as audio amplifiers, but they are used as r-f power amplifiers in transmitters. If power is delivered to a tuned load the load will present a

high impedance at the resonant frequency and low impedance at other frequencies. If the load is tuned to the same frequency as that which is applied to the grid it will offer optimum loading at this frequency. Low impedance will be offered to the harmonics (multiples) of the frequency applied to the grid, and hence these undesirable components will be eliminated.

According to Frequency

Amplifiers may be classified according to the frequency range over which they are to operate. In general, amplifiers operating in these ranges are known as direct-current (d-c); audio-frequency (a-f); intermediate-frequency (i-f); radio-frequency (r-f); and video-frequency (v-f), or pulse, amplifiers.

When the signal current is in but one direction a d-c amplifier must be used. In order to overcome certain problems inherent in such an amplifier the circuits must be balanced and stabilized by means of resistors.

Audio-frequency amplifiers operating in the range from 30 to 15,000 cycles per second may be transformer-coupled, impedance-coupled, or resistance-coupled.

I-f and r-f amplifiers are ordinarily designed for tuned-circuit coupling, although in actual operation they may resemble either transformer-coupled or impedance-coupled circuits.

Video-frequency amplifiers, which operate in a range extending from the lower audio frequencies to perhaps 5,000,000 cycles per second, commonly use resistance-coupled amplifiers in which the coupling resistance is made low enough to produce the necessary high-frequency response. However, in actual radar and television applications the resistance-coupled amplifier must be modified to make the response essentially flat over a wide range of frequencies. In addition, the circuits must be modified to keep time-delay distortion within a certain minimum value at the high- and low-frequency ends of the spectrum.

According to Circuit Configuration

GROUNDED-CATHODE AMPLIFIER.—Amplifiers may be classified according to the connection of the tube elements in the circuit. Conventional electron-tube amplifier circuits return the cathode either to ground through a cathode resistor or, if separate bias is provided, to ground directly. Both of these circuit configurations are classed as grounded-cathode

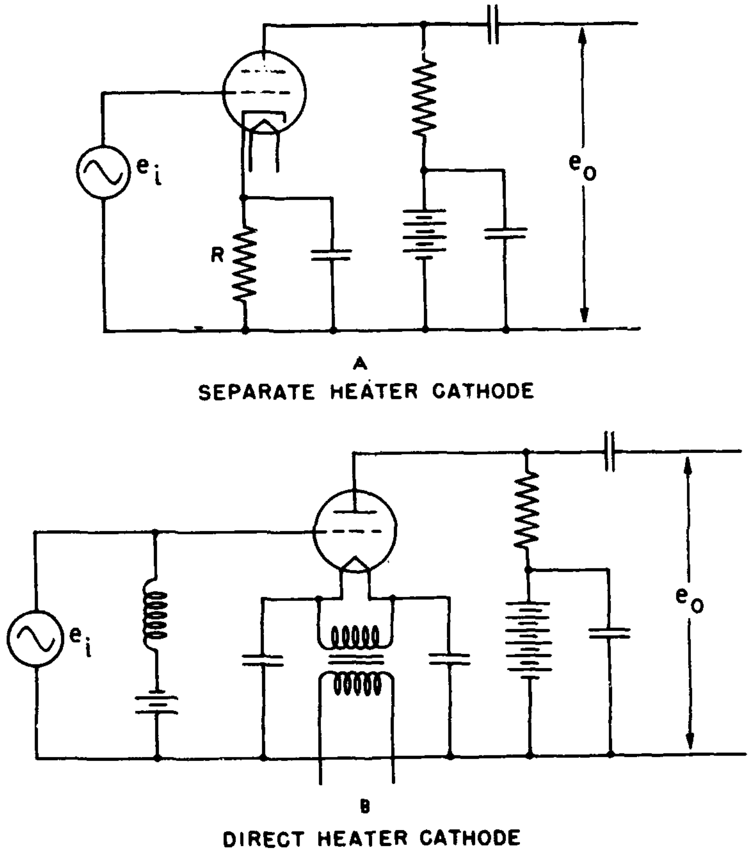


Figure 4-7.—Grounded-cathode amplifier circuits.

types. In the former, the cathode is positive with respect to ground by the amount of the grid bias and the cathode bypass capacitor holds the cathode at ground potential with respect to the signal component. A grounded-cathode amplifier circuit is shown in figure 4-7, A, for a separate heater cathode and in figure 4-7, B, for a direct heater cathode. In both of these circuits the interelectrode capacitance between plate and grid introduces feedback at high frequencies, and unstable amplifier operation results.

GROUNDED-GRID AMPLIFIER.—For very high frequencies the grounded-grid amplifier shown in figure 4-8 removes the

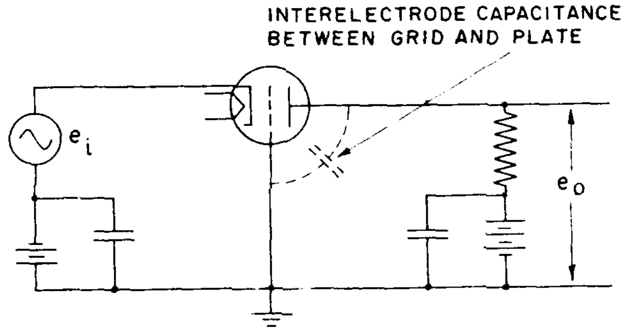


Figure 4-8.—Grounded-grid amplifier circuit.

feedback coupling between grid and plate and places the grid-plate interelectrode capacitance effectively in parallel with the load. Grounded-grid amplifiers are used as r-f amplifiers in the lower radar frequencies and in television circuits in the v-h-f and u-h-f bands. The input signal is introduced into the cathode circuit in series with the grid bias and varies the grid-to-cathode volt[illegible] in the normal manner.

The output signal is taken between the plate and ground. The plate current (including the a-c component) flows through the signal source which is in series with the cathode circuit. The signal source has appreciable impedance and

the plate current through it is accompanied by a voltage drop across it which acts between the cathode and grid. The action is degenerative and lowers the gain of the amplifier compared to the gain of the grounded-cathode type. Some of the power in the load is supplied by the signal source since the load and source are in series with the plate-to-cathode resistance of the tube. The source is thus required to furnish a considerable amount of power.

GROUNDED-PLATE AMPLIFIER.—The grounded-plate amplifier (*V*1 in fig. 4-9) is another circuit configuration that

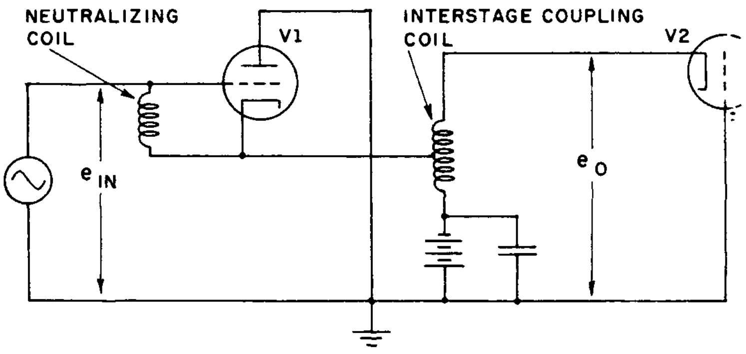


Figure 4-9.—Grounded-plate amplifier.

may be used in u-h-f amplifiers. This amplifier circuit has a lower signal-to-noise ratio than the grounded-cathode grounded-grid circuits and also poorer stability since the grid is not used as a shield between cathode and plate. However, the input capacitance is slightly less than that of the grounded-grid circuit. If the grounded-plate amplifier is used to drive a grounded-grid amplifier, *V*2, at very high frequencies, the lower induced grid noise gives this circuit configuration a slight advantage over that of the grounded-cathode grounded-grid type. Generally, however, the grounded-cathode grounded-grid circuit is preferred.

According to Resonant Quality of Load

Amplifiers are also classified according to whether they are TUNED or UNTUNED—that is, according to whether they amplify a restricted range or a wide range of frequencies respectively.

TUNED AMPLIFIERS.—Tuned amplifiers may be further subdivided into NARROW-BAND and WIDE-BAND amplifiers. Whether a band of frequencies is considered narrow or wide depends on the ratio of the bandwidth to the center frequency, expressed as a percentage of the center frequency. An example of a narrow-band amplifier is the i-f amplifier in a broadcast radio receiver. The range is about 10 kc with a center frequency of 450 kc. The bandwidth in this example is 2.2 percent of the center frequency. Examples of wider-band amplifiers are the i-f stages of radar and television receivers which may have a range of about 4 mc at a center frequency of about 30 mc. In this case the bandwidth is about 14 percent of the center frequency.

UNTUNED AMPLIFIERS.—Untuned amplifiers are not tuned to any specific band of frequencies. The circuit components, however, may limit the range of frequencies that the circuit can handle. All audio amplifiers come under this classification.

DISTORTION IN AMPLIFIERS

The output of an ideal amplifier is identical with the input in all respects except for an increase in amplitude. This statement precludes, of course, the various wave-shaping and special-purpose amplifiers. A practical amplifier, however, falls short of this ideal. Not all frequency components present in the input may be amplified equally; the amplitude of the output voltage may not be proportional to the amplitude of the input voltage, and thus new frequencies will be introduced; or the relative phases of the various output frequencies may differ from those of the input. These deviations from the ideal are known as FREQUENCY DISTOR-

TION, AMPLITUDE (or nonlinear) DISTORTION, and PHASE (or delay) DISTORTION, respectively.

To achieve the special waveforms necessary in certain radar, television, or test circuits distortion is deliberately introduced by the amplifier or an associated circuit. In certain other circuits, however, less distortion of all three types is permitted than would be tolerated in the case of broadcast radio amplifiers.

Frequency Distortion

When some frequency components of a signal are amplified more than others or when some frequencies are excluded, the result is frequency distortion. Essentially, this type of distortion results from bandwidth restrictions imposed by the various amplifier circuit components. For example, if a coupling circuit does not pass the third or higher harmonics that are present in the input, the circuit introduces frequency distortion.

For purposes of comparison figure 4–10, A, shows the input and output of a two-stage amplifier that has introduced frequency distortion. The input, e_{in}, contains the fundamental and the third harmonic; but the output, e_{out}, contains only the fundamental since the amplifier is unable to pass the third harmonic. Frequency distortion may occur at low frequencies if the coupling capacitor between the stages is so small that it presents a high series impedance to the low-frequency components of a signal. Distortion may also occur at high frequencies because of the shunting effects of the distributed capacitance in the circuit.

Low- and high-frequency compensation—that is, boosting the response of the amplifier at the low- and high-frequency ends of the desired band—is discussed in chapter 5 under "Video Amplifiers."

Phase Distortion

Most coupling circuits shift the phase of a sine wave, but this shift has no effect on the shape of the output. How-

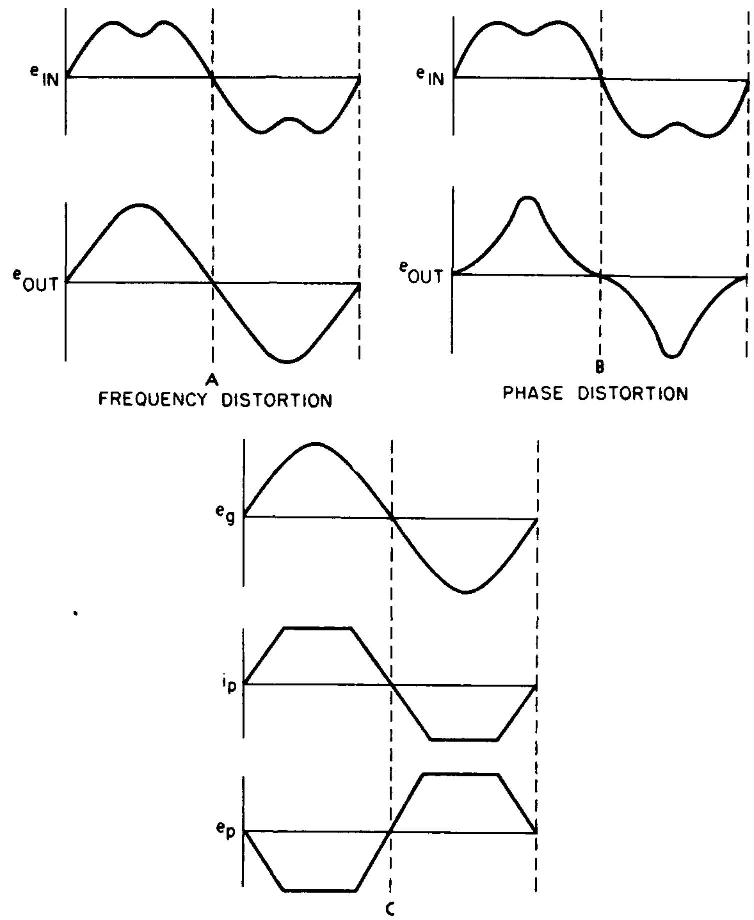


Figure 4-10.—Types of amplifier distortion.

ever, when more complex waveforms are amplified, each component frequency that makes up the over-all waveform may have its phase shifted by an amount that depends on its frequency. Thus, the output is not a faithful reproduction of the input waveform.

Figure 4–10 B, shows the input and output waveforms of a two-stage amplifier that has introduced phase distortion. The input signal, e_{in}, consists of a fundamental and a third harmonic. Although the amplitudes of both components have been increased by identical ratios, the output, e_{out}, is considerably different from the input because the phase of the third harmonic has been shifted with respect to the fundamental.

Basically, phase distortion is present whenever the component frequencies in the input of an amplifier are not all passed through the amplifier in the same amount of time. Phase or time-delay distortion is not important in the amplification or reproduction of sound because the ear is unable to detect relative phase shifts of the individual components. Such distortion, however, is important in radar, television, and measuring equipment where the waveform must be accurately maintained during amplification. Phase distortion may be reduced by varying the amount or type of coupling. In video amplifiers special coupling circuits are used to reduce this distortion.

Amplitude Distortion

If a signal is amplified by an electron tube that is not operating on the linear portion of its characteristic curve, amplitude (nonlinear) distortion will occur. In the nonlinear region a change in grid voltage does not result in a change in plate current that is directly proportional to the change in grid voltage. For example, if a tube is overdriven by applying a grid signal that drives the tube beyond the linear portion of the characteristic curve (nonlinear distortion) and also to the point where the grid draws current (grid-limiting distortion) the resultant signal is distorted in amplitude, as shown in figures 4–10, C, and 4–11. This t pe of distortion is to be expected, since for a portion of the negative half of the grid signal swing the tube operates on a nonlinear portion of the characteristic curve, and for a portion of the positive swing the grid draws current.

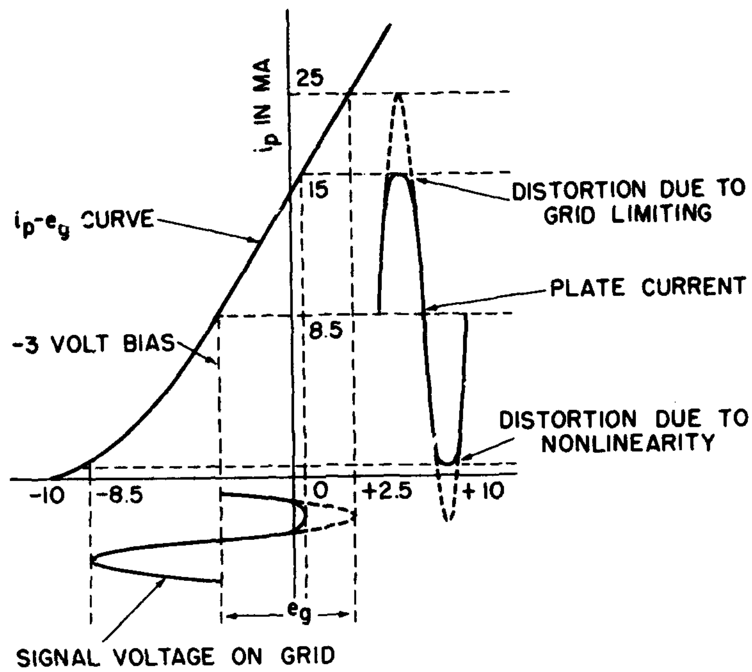


Figure 4-11.—Distortion in a class-A amplifier due to excessive signal voltage.

Beyond the linear portion of the curve, a further increase in negative grid potential will not cause a proportionate reduction in plate current. On the other half cycle, the positive swing of grid voltage (beyond the point where the grid draws current) is limited by the loss in voltage within the source impedance and no further increase in plate current can occur. The result of this nonlinearity is the production of harmonics that were not present in the input of the amplifier. This concept can be better understood if it is recalled that any complex periodic waveform may be considered as being composed of a number of sine waves of different frequencies and amplitudes. The sine wave that has the same frequency as the complex periodic wave is

called the FUNDAMENTAL. The frequencies higher than the fundamental are called HARMONICS. Thus, from the complex waveform is obtained a number of harmonics plus the fundamental frequency.

At the higher frequencies, harmonics may be reduced by the use of a parallel resonant circuit as a plate load, by link coupling, or by filtering. At the audio frequency, however, there is an overlap of frequencies and filtering is not practicable. The best solution is to operate the tube on the straight portion of the characteristic curve for class-A operation, or to operate it in a push-pull arrangement for class-B operation.

Complex waveforms are necessary in certain television, radar, and test instrument circuits. These waveforms include square waves, saw-tooth waves, and peaked waves. In each of these waveforms the distortions are deliberately introduced. The circuits for producing these distorted waveforms and an analysis of nonsinusoidal waves and transients are treated in an advanced course.

Miscellaneous Distortion

HUM is a type of distortion particularly objectionable in audio- and video-frequency amplifiers. It may be caused by one or more of the following conditions: Alternating current in the filaments or heaters of the amplifier tubes, stray electromagnetic or electrostatic fields, or insufficient filtering of the power supply. A center-tapped resistor across the filament terminals to which the grid return is connected may reduce hum at the power frequency. On the other hand, the elimination of hum due to cyclic variations in filament temperature, at twice the power frequency, is largely a design problem. The elimination of hum in heater-type tubes is also largely a design problem, although, as in the case of filament-type cathodes, the a-c leads may be twisted together and placed in such positions as to cause the least magnitude of induced voltages in the signal circuits.

The undesired effects produced by stray fields may be

reduced by proper placement of transformers; proper shielding of transformers, leads, and tubes; and arrangement of circuits and components so that there will always be a low impedance bypass to ground to the undesired currents.

MICROPHONIC EFFECTS are the result of slight vibrations in the tube elements. Variations in plate current due to these vibrations are amplified in each succeeding stage and appear in the output of audio and video amplifiers. These slight displacements of the tube elements may be caused either by physical vibration of the chassis or by the sound vibrations emitted by the speaker.

The obvious remedy is to employ some method that will insulate the tube or tubes from the vibrating source. Some tubes, however, are less susceptible to microphonics than others, and occasionally simply replacing a tube will cure the trouble.

NOISE in audio and video amplifiers may be caused by faulty contacts, faulty components such as resistors or capacitors, or THERMAL-AGITATION NOISE. Thermal-agitation noise occurs because all electrical conductors contain electrons moving at random. Some of these electrons move at random even if there is an impressed voltage across the conductor. By chance, at any given instant, more of these electrons move in one direction than in another. When amplified, the accompanying voltage results in thermal-agitation noise.

Also inherent in electron tubes are other noises such as SHOT EFFECT, which results from a variation in the rate of electron emission from a cathode; GAS NOISE, which results from a variation in the rate of production of ions; and SECONDARY EMISSION NOISE, which results from a variation in the rate of production of secondary electrons. There are also other variations that produce noise in the output of a receiver.

In the final analysis, tube noise is the limiting factor that determines the ultimate sensitivity of an amplifier.

COUPLING METHODS

A single stage of voltage or power amplification normally is not sufficient for radio or radar applications. To obtain the necessary gain, several stages must often be connected together. The output of one stage then becomes the input of the next throughout the series of stages, and this arrangement is called a CASCADE AMPLIFIER.

A cascade amplifier is designated according to the method used to couple one amplifier stage to the next. There are a number of methods, each having certain advantages and disadvantages, and the choice for a particular application depends on the needs of the circuit. The basic methods are: (1) resistance-capacitance coupling, (2) impedance coupling, (3) transformer coupling, and (4) direct coupling.

Before considering the details of each coupling method it is desirable to establish the equivalent circuit of an electron-tube amplifier. The characteristics of an amplifier are determined more readily by replacing the tube with its equivalent circuit and analyzing this circuit.

Equivalent Circuit of an Electron-Tube Amplifier

The analysis of the electron-tube amplifier (fig. 4–12, A) is facilitated by the use of an equivalent circuit in which the a-c components of current and voltage are present and the d-c components are not. The waveforms of current and voltage in the actual circuit are shown in figure 4–12, B. The equivalent circuit may be either of two forms.

In the CONSTANT-VOLTAGE GENERATOR FORM of equivalent circuit (fig. 4–12, C), the tube is replaced by a generator acting in the plate-cathode circuit that develops a voltage, $-\mu e_s$. The generator has an internal resistance equal to the plate resistance of the tube and is connected in series with the load impedance.

In the CONSTANT-CURRENT GENERATOR FORM of equivalent circuit (fig. 4–12, D), the tube is replaced by a generator acting in the plate-cathode circuit that develops a current,

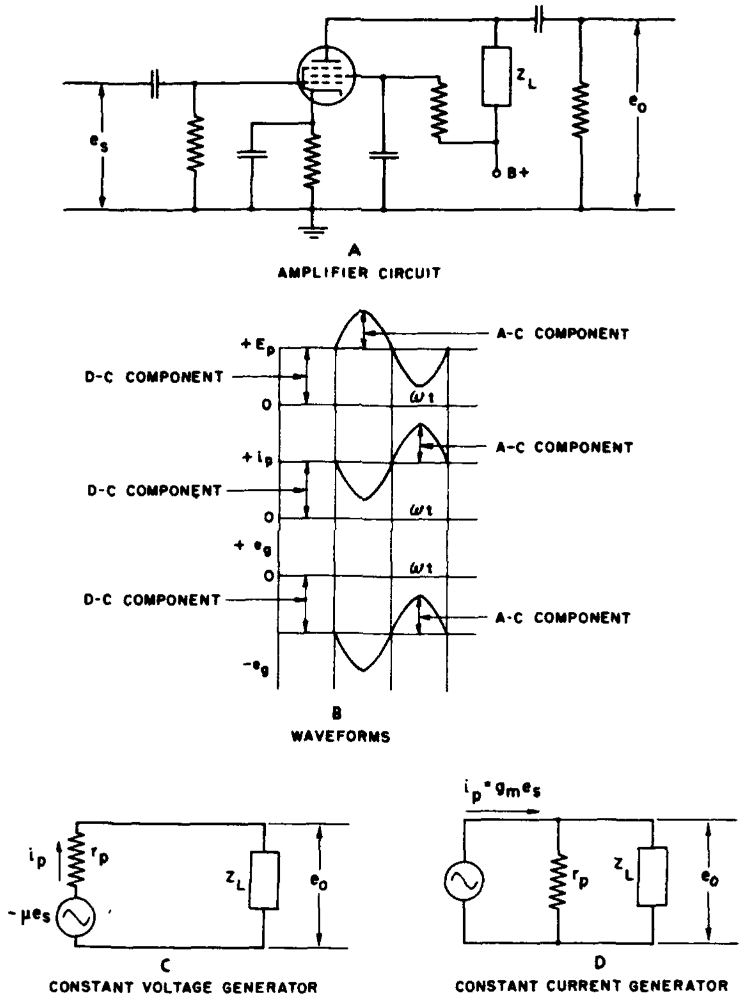


Figure 4-12.—Simplified amplifier circuits and waveforms.

$-g_m e_s$. Again, the generator resistance is equal to the tube plate resistance but in this circuit the load impedance acts in parallel with the generator resistance, not in series with it. The constant-current generator form is more convenient for pentodes in which the plate resistance is much higher than the load impedance.

The minus signs used with the expressions for voltage and current in the equivalent circuits indicate merely that these signals are of opposite polarity to those of the grid since these signals act in the plate-cathode circuit, not in the grid circuit. It will be recalled that when the a-c signal applied to the grid swings in a negative direction (fig. 4–12, B), plate current decreases and the voltage drop across the plate load impedance decreases. Since the B-supply voltage is constant, the plate voltage increases (swings in a positive direction). Thus the grid signal and the plate signal are of opposite instantaneous polarity, or 180° out of phase.

In figure 4–12, C, the a-c component of plate current is

$$i_p = \frac{-\mu e_s}{r_p + Z_L}.$$

The a-c component of output voltage appears across Z_L as

$$\begin{aligned} e_o &= i_p Z_L \\ &= \frac{-\mu e_s Z_L}{r_p + Z_L}. \end{aligned}$$

It is thus apparent that the output voltage of an amplifier is not simply μ times the applied signal. A part of the total voltage acting in the equivalent circuit of figure 4–12, C, is developed across the internal resistance of the generator and is thus not available across the load impedance. The load impedance and plate resistance are in effect a voltage divider across which the total a-c voltage generated within the tube is applied.

In the constant-current generator form of equivalent circuit (fig. 4–12, D), the output voltage is the voltage ap-

pearing across the load impedance in parallel with the plate resistance. The output voltage is

$$e_o = \frac{i_p r_p Z_L}{r_p + Z_L}$$

$$= \frac{-g_m e_s r_p Z_L}{r_p + Z_L}.$$

Since $g_m = \frac{\mu}{r_p}$,

$$e_o = \frac{-\mu e_s r_p Z_L}{r_p(r_p + Z_L)}$$

$$= \frac{-\mu e_s Z_L}{r_p + Z_L}.$$

Thus the same output voltage is obtained in both constant-voltage and constant-current generator forms of equivalent circuits.

The equivalent circuit gives the exact performance of the amplifier *only to the extent that* r_p and μ (used in the equivalent circuits) are constant over the range of variations produced in the control-grid and plate voltages by the signal voltage. Hence, when the signal voltage is small the equivalent circuit is almost exactly correct; but if the signal voltage is increased, the error involved in the equivalent circuit becomes proportionately larger. If the exact behavior of the amplifier is to be ascertained, the equivalent circuit must be modified to take into account the effects produced by variations in circuit constants.

Resistance-Capacitance Coupling

One of the most widely used methods of connecting amplifier stages is *R-C* coupling. Amplifiers coupled in this manner are relatively inexpensive, lack heavy components, have good fidelity over a comparatively wide frequency range, are relatively free from undesirable induced currents from a-c heater leads, and are especially suitable for use with pentodes and high-mu triodes.

A resistance-capacitance coupled amplifier (generally shortened to resistance-coupled amplifier) can be designed to have good response for almost any desired frequency range. For instance, it can be designed to give fairly uniform amplification of all frequencies in the range from 100 to 20,000 cps.

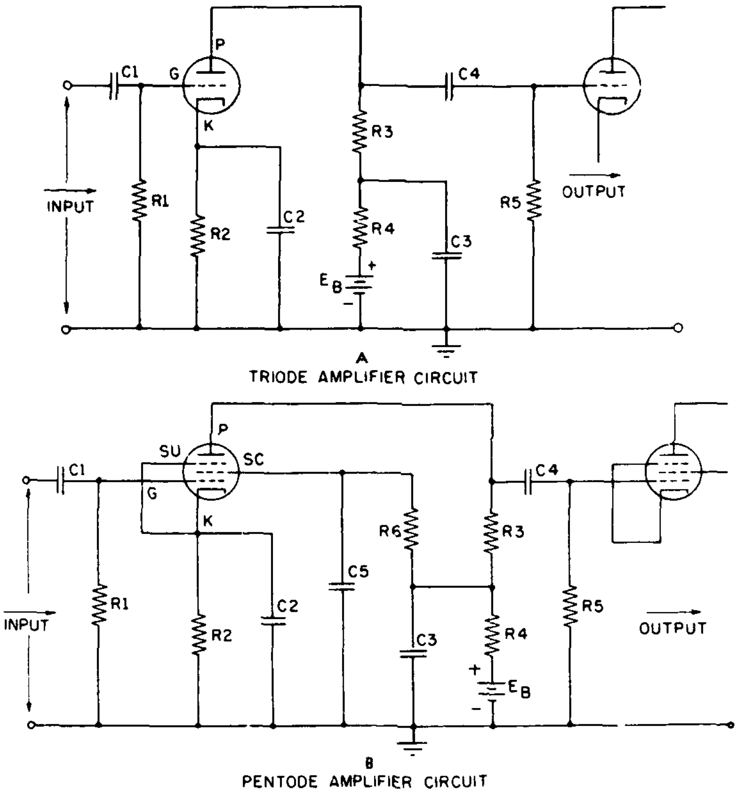


Figure 4-13.—Typical resistance-coupled amplifiers.

R1—Grid-leak resistor.
R2—Cathode bias resistor.
R3—Plate load resistor.
R4—Plate decoupling resistor.
R5—Second-stage grid resistor.
R6—Screen dropping resistor.

C1—Input coupling.
C2—Cathode bypass capacitor.
C3—Plate supply bypass capacitor.
C4—Output coupling capacitor.
C5—Screen bypass capacitor.

Slight modification of the circuits can extend the frequency to cover the wide band required in video amplifiers. However, extension of the range can be obtained only at the cost of reduced amplification over the entire range. Thus the R-C method of coupling amplifiers gives a good frequency response with minimum distortion, but it also gives low amplification.

Typical resistance-coupled amplifiers are shown in figure 4-13, together with the names of the various circuit elements.

In the triode shown in figure 4–13, A, the d-c grid circuit includes *G*, *R*1, *R*2, and *K*; and the a-c grid circuit includes *G*, *R*1, *C*2, and *K*. In the pentode in figure 4–13, B, the d-c screen circuit includes *SC*, *R*6, *R*4, E_B, *R*2, and *K*; and the a-c circuit includes *SC*, *C*5, *C*2, and *K*. In each case the d-c plate circuit includes *P*, *R*3, *R*4, E_B, *R*2, and *K*; and the a-c plate circuit includes *P*, *R*3, *C*3, *C*2, and *K*.

In order that the output voltage may be large, the load resistor should have as high a value as practicable. However, the higher this value becomes, the greater is the voltage drop across it and the lower is the voltage remaining between the plate and cathode of the tube. To obtain the required effective plate voltage, the voltage drop across the load resistor is subtracted from the plate supply voltage. Thus there is a practical limit to the size of the plate load resistor if the plate is to be supplied with its rated voltage. If a larger plate resistor is necessary, the only alternative is to increase the plate supply voltage. There is, of course, a practical limit to the amount that the plate voltage may be increased. An example of the d-c voltage distribution around the plate circuit is shown in figure 4–14. The plate current is 6 ma and the voltage across the 30 k-ohm load resistor is 6×30, or 180 volts. The voltage drop across the 500-ohm cathode resistor is 0.006×500, or 3 volts, which provides the grid bias for the tube. The plate voltage is the B-supply voltage less the drop through R_L and R_K, or $300-180-3=117$ volts.

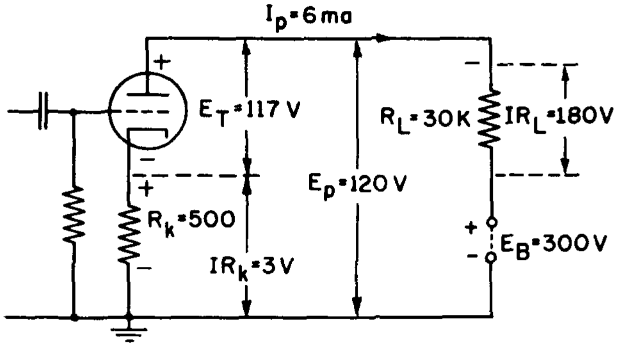


Figure 4-14.—Plate-circuit voltage distribution.

The screen resistor, *R*6, in figure 4–13, B, has the necessary voltage drop across it so that when this drop is subtracted from the B-supply voltage, the rated screen voltage will remain. The value of the cathode resistor, *R*2, is determined by the grid bias required and the no-signal plate current.

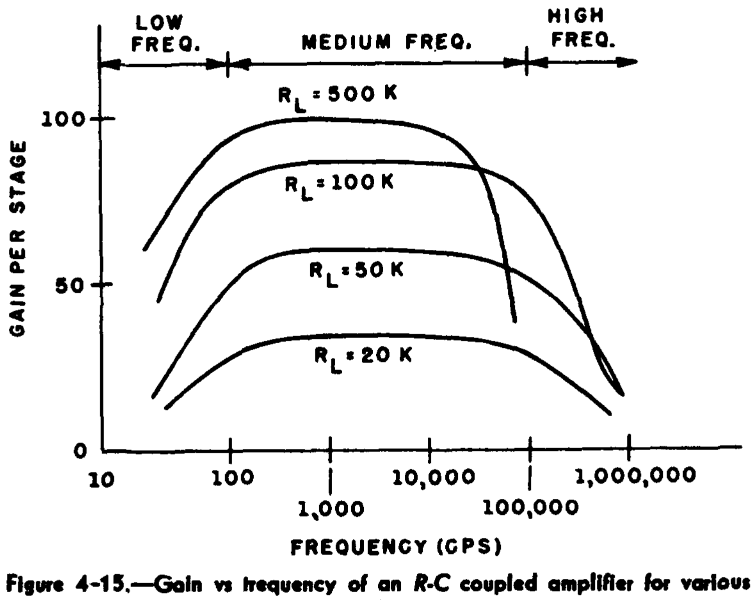


Figure 4-15.—Gain vs frequency of an *R-C* coupled amplifier for various plate loads.

For the range of frequencies to be amplified, the cathode bypass capacitor, $C2$, has a low reactance to the a-c component of plate current in comparison with the resistance of $R2$. The decoupling (or filter) circuit, $C3R4$, tends to prevent the a-c component of plate current from flowing through the B supply because $R4$ offers a high series resistance and $C3$ offers a low shunt reactance to the a-c signal component.

Typical frequency response curves for an R-C coupled audio amplifier are shown in figure 4–15. The response is measured in terms of the voltage gain of the amplifier over a range of frequencies. The voltage gain is the ratio of e_o to e_s. The gain falls off at very low frequencies because of the increase in the capacitive reactance of the interstage coupling capacitor, C_c (fig. 4–16, A). This capacitor acts in series between the source and the load and has developed across it an increasing part of the signal voltage as the frequency is decreased.

The reduction in gain at the higher frequencies is due to the fact that the load resistor (R_L) is shunted by the output capacitance (C_o) of one stage, the input capacitance (C_i) of the next stage, and the distributed capacitance (C_d) of the coupling network. The combined effect of these capacitances is to increase the part of the total signal voltage that is developed across the internal resistance, r_p, of the generator (fig. 4–16, B) and to decrease the part that appears as output voltage, e_{out}.

Middle-frequency gain.—The middle-frequency gain is flat and in the example in figure 4–15 is assumed to extend approximately from 100 to 200,000 cps. The equivalent circuits shown in figure 4–17 represent the active circuit components and their connections for the middle-frequency range. Figure 4–17, A, illustrates the constant-voltage generator form and figure 4–17, B, the constant-current generator form. The reactance of the coupling capacitor, C_c, is low at the middle frequencies and thus is omitted between R_L and R_{g1}. The reactances of the shunting capacitances, C_o, C_i, and C_d, are high at the middle frequencies

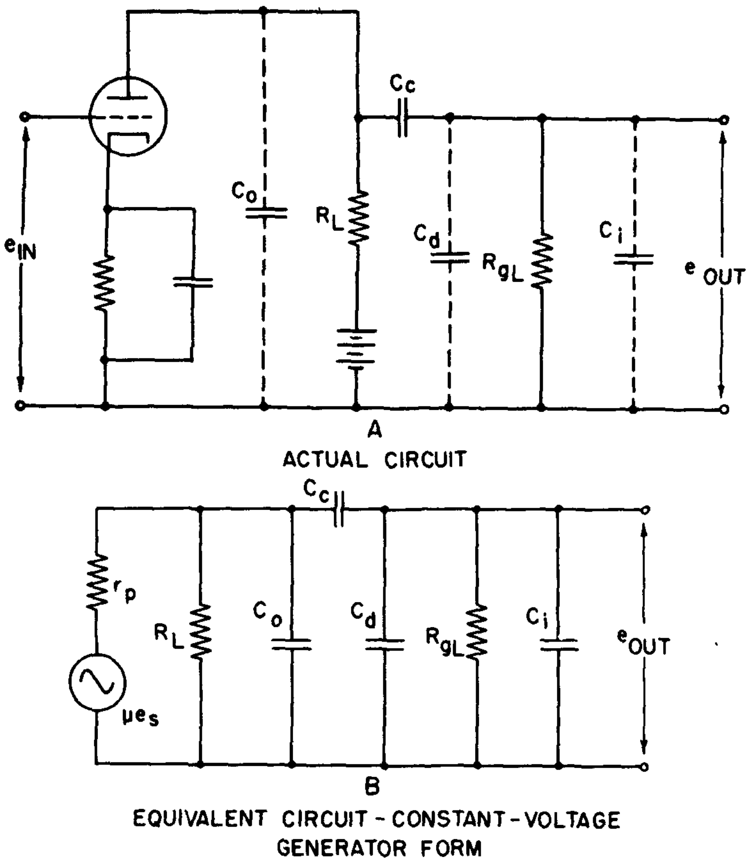
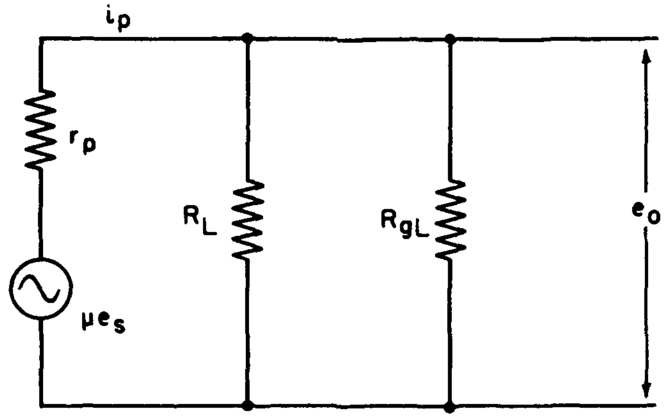


Figure 4–16.—Single-stage resistance-coupled amplifier and equivalent circuit.

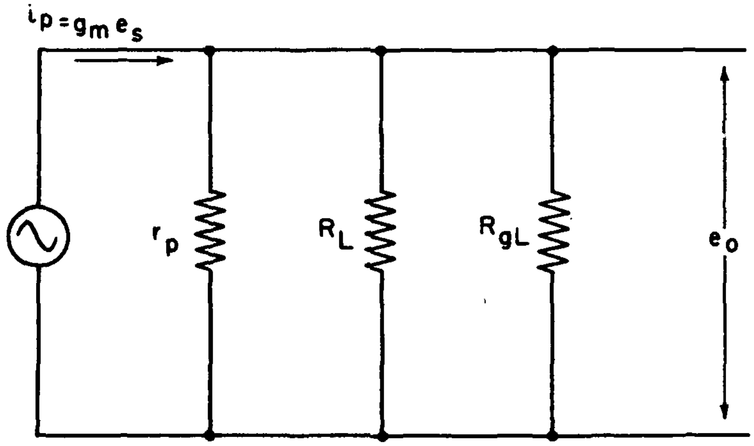
and so they too are omitted from the equivalent circuit. Thus the equivalent circuits are reduced to include only the generator, r_p, R_L, and R_{gl}. The amplification is independent of frequency, and a flat response may be expected.

In the constant-voltage generator form (fig. 4–17, A), the a-c component of plate current is

$$i_p = \frac{\mu e_s}{r_p + R_{eq}},$$



A
CONSTANT-VOLTAGE FORM



B
CONSTANT-CURRENT FORM

Figure 4-17.—Middle-frequency equivalent circuits.

where R_{eq} is the combined resistance of R_L and R_{gl} in parallel. The output voltage across R_{eq} is

$$e_o = i_p R_{eq}$$

$$= \frac{\mu e_s R_{eq}}{r_p + R_{eq}}.$$

The voltage gain is

$$\frac{e_o}{e_s} = \frac{\mu R_{eq}}{r_p + R_{eq}}.$$

For example, if the triode has an amplification factor of 20, a plate resistance of 10 k-ohms, a plate load resistance of 50 k-ohms, and a grid-leak resistance of 100 k-ohms, the middle-frequency voltage gain is

$$\frac{e_o}{e_s} = \frac{20 \times \frac{50 \times 100}{50 + 100}}{10 + \frac{50 \times 100}{50 + 100}} = \frac{20 \times 33.3}{10 + 33.3} = 15.4.$$

In the constant-current generator form (fig. 4–16, B), the output voltage appearing across r_p, R_L, and R_{gl} in parallel is

$$e_o = i_p \frac{1}{\frac{1}{r_p} + \frac{1}{R_L} + \frac{1}{R_{gl}}}$$

$$= g_m e_s \frac{1}{\frac{1}{r_p} + \frac{1}{R_L} + \frac{1}{R_{gl}}}.$$

The voltage gain is

$$\frac{e_o}{e_s} = g_m \frac{1}{\frac{1}{r_p} + \frac{1}{R_L} + \frac{1}{R_{gl}}}.$$

For example, if an *R-C* coupled pentode has a plate resistance of 1 megohm, a plate-load resistance of 0.125 megohm, a grid-leak resistance of 0.25 megohm, and a transconductance of 900 micromhos, the middle-frequency gain is

$$\frac{e_o}{e_s}=900\times10^{-6}\times\frac{1}{\frac{1}{1\times10^6}+\frac{1}{0.125\times10^6}+\frac{1}{0.25\times10^6}}$$

$$=900\times10^{-6}\times\frac{10^6}{1+8+4}$$

$$=69.4$$

The voltage gain of 69.4 is less than 10 percent of the amplification factor of the pentode because the combined resistance across which the output voltage appears is less than 10 percent of the pentode plate resistance.

LOW-FREQUENCY LIMIT.—In the lower range of frequencies amplified by an *R-C* coupled amplifier, C_o, C_d, C_i, are unimportant and are omitted from the equivalent circuit since the X_C ohms are high and are in parallel with R_L and R_{gl}. The reactance of coupling capacitor C_c, however, becomes increasingly important at the low frequencies and cannot be neglected. A low-frequency equivalent circuit is shown in figure 4–18.

Since X_C varies inversely with frequency more of the total voltage, μe_s, appears across C_c and less across R_{gl} as the frequency decreases. The approximate frequency at which the output voltage falls to 70 percent of its value at the

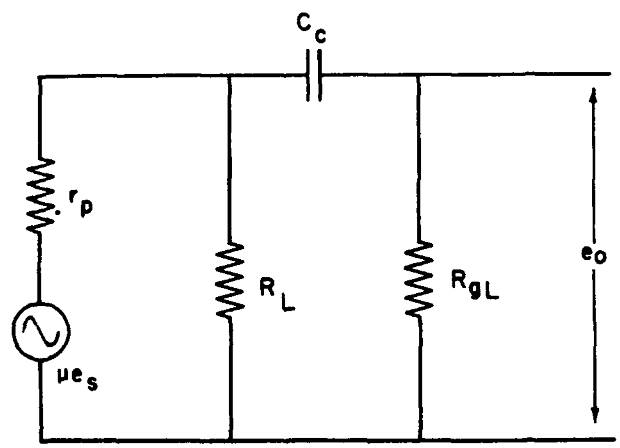


Figure 4–18.—Low-frequency equivalent circuit.

middle frequencies is the frequency at which X_C of the coupling capacitor is equal to R_{gl}. The approximation is applicable to triode amplifiers. The low-frequency limit is derived as follows,

$$X_C = R_{gl}$$

$$\frac{1}{2\pi f C_c} = R_{gl}$$

$$f = \frac{1}{2\pi C_c R_{gl}}.$$

For example, an R-C coupled triode amplifier has a coupling capacitor of 0.04 μf and a grid-leak resistor of 100,000 ohms. The low-frequency limit is

$$f = \frac{1}{2\pi \times 0.04 \times 10^{-6} \times 10^{5}}$$

$$= 39.8 \text{ cps.}$$

The low-frequency limit of approximately 40 cps is the frequency at which the output voltage is 70 percent of its middle-frequency value.

Increasing the size of C_c lowers the frequency response of the amplifier but there is a practical limit. This limit is caused by a type of regeneration that may occur between several R-C coupled stages supplied by a common plate and screen power source. Such regeneration is called MOTOR BOATING and occurs when the B-supply source impedance is relatively high compared with the X_C ohms of the interstage coupling capacitors.

HIGH-FREQUENCY LIMIT.—In the high-frequency range the shunting capacitances, C_o, C_d, and C_i, of the general equivalent circuit (fig. 4–16, B) become significant. These capacitances limit the output voltage at high frequencies. In figure 4–19 these parallel capacitances are combined and designated "C_s." The high-frequency limit is the frequency at which the output voltage falls to 70 percent of its value

at the middle frequencies. This limit occurs at the frequency at which the X_C ohms of the shunting capacitance, C_s, are equal to the combined resistance of r_p, R_L, and R_{gl} in parallel. Thus

$$\frac{1}{2\pi f C_s}=R_{eq},$$

and

$$f=\frac{1}{2\pi C_s R_{eq}}.$$

For example, an R-C coupled pentode amplifier has a plate resistance of 1 megohm, a load resistance of 0.125 megohm, a grid-leak resistance of 0.25 megohm, and a shunting capacitance of 100 $\mu\mu$f. The combined resistance of r_p, R_L, and R_{gl} in parallel is

$$R_{eq}=\frac{1}{\frac{1}{1}+\frac{1}{0.125}+\frac{1}{0.25}}$$

$$=\frac{1}{1+8+4}$$

$$=\frac{1}{13}=0.077 \text{ megohm}.$$

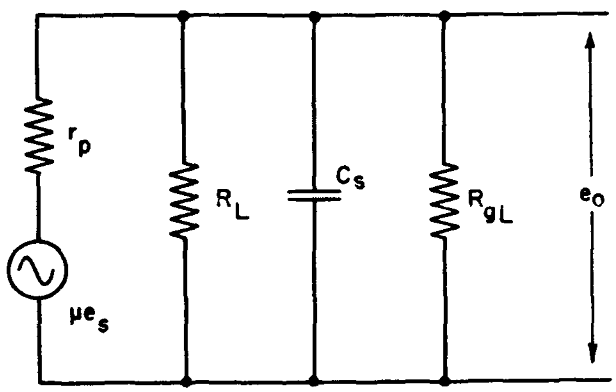


Figure 4-19.—High-frequency equivalent circuit.

The high-frequency limit is

$$f=\frac{1}{6.28\times100\times10^{-12}\times0.077\times10^{6}}$$

$$=20{,}700 \text{ cps.}$$

The high-frequency limit of approximately 21,000 cps is the frequency at which the output voltage falls to 70 percent of its middle-frequency value. The upper-frequency limit may be extended by using tubes having low interelectrode capacitances. The upper limit may also be extended by reducing R_L, but at the expense of mid-frequency gain.

Wide-band *R-C* coupled amplifiers are characterized by many stages having relatively low gain per stage, large coupling capacitors between stages, and low-shunting capacitances.

The coupling capacitor must have a high d-c resistance (low leakage) because it insulates the grid circuit of the output stage from the plate voltage of the driver stage. The d-c resistance of a coupling capacitor should be at least 50 megohms. If the leakage becomes appreciable, direct current flows through R_{gl} and the resulting voltage impresses a positive bias on the next tube. This condition is especially important if an *R-C* coupled amplifier has a very good low-frequency response because in such a case both the grid-leak resistor and coupling capacitor are large.

Impedance Coupling

Impedance or inductance-capacitance coupling is obtained by replacing the load resistor, R_L, of a normal *R-C* coupled amplifier with an inductance, L, as shown in figure 4–20. To obtain as much amplification as possible, particularly at the lower frequencies, the inductance is made as large as practicable. To avoid undesirable magnetic coupling a closed-shell type of inductor is used. Because of the low d-c resistance of the inductor, less d-c voltage appears across it. Thus the tube can operate at a higher plate voltage.

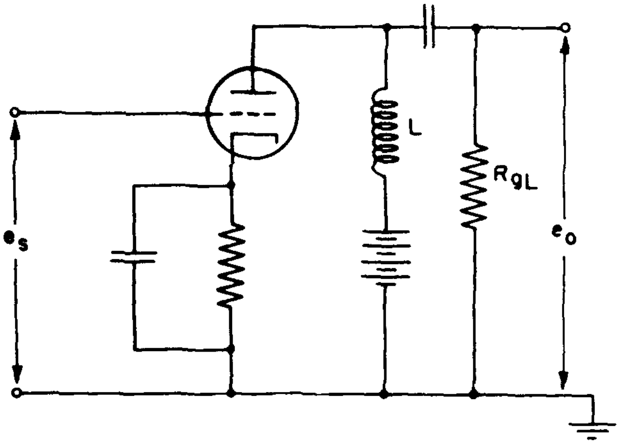


Figure 4–20.—Impedance-coupled amplifier.

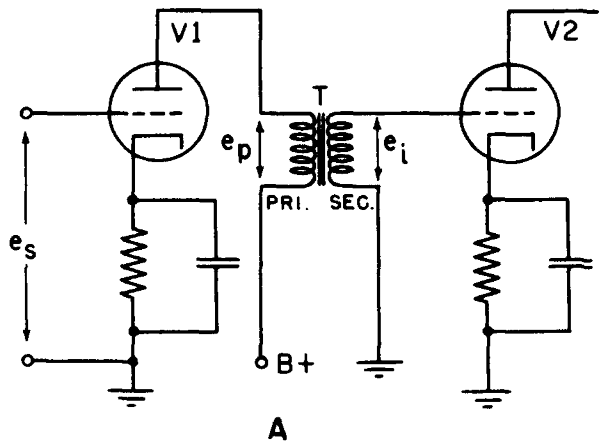
The degree of amplification is not uniform as it is with the *R-C* coupled amplifier because the load impedance, Z_L, varies with the frequency—that is,

$$Z_L = R + j2\pi fL.$$

Since the output voltage appears across Z_L, the voltage gain increases with the frequency up to the point where the shunting capacitance limits it. The shunting capacitance includes not only the interelectrode and distributed wiring capacitances found in *R-C* coupled amplifiers but also the distributed capacitance associated with the turns of the inductor. The distributed capacitance between the turns of the coil greatly increases the capacitance to ground and plays a major part in limiting the use of this coupling at the higher frequencies.

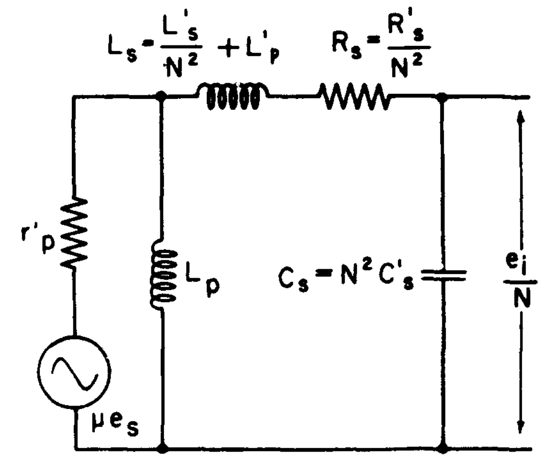
Transformer Coupling

A transformer-coupled stage of amplification (fig. 4–21) has certain advantages over other types of coupling. The voltage amplification of the stage may exceed the amplification of the tube if the transformer has a step-up turns ratio. Direct-current isolation of the grid of the next tube is pro-



A
ACTUAL CIRCUIT

- e_s—input signal.
- e_p—primary signal voltage.
- e_i—input voltage to second tube.
- r_p—plate resistance of tube & resistance of primary windings.
- μ—Amplification factor of tube.
- L_p—incremental inductance of primary.
- L'_p—primary leakage inductance.



B
EQUIVALENT CIRCUIT REDUCED TO UNITY TURNS RATIO

- L'_s—secondary leakage inductance.
- R'_s—resistance of secondary windings.
- C'_s—distributed capacitances of secondary circuit.
- N—secondary-to-primary turns ratio of transformer. (unity turns ratio assumed in the equivalent circuit).

Figure 4-21.—Single-stage transformer-coupled voltage amplifier and equivalent circuit.

vided without the need for a blocking capacitor; and the d-c voltage drop across the coupling resistor, which is necessary when *R-C* coupling is used, is avoided. This type of coupling is also used to couple a high-impedance source to a low-impedance load, or vice versa. Also it may be used as a simple means of providing phase inversion for a push-pull amplifier without the use of special phase-inverting circuits.

Transformer coupling has the disadvantages of greater cost, greater space requirement, the necessity for greater shielding, and the possibility of poorer frequency response at the higher and lower frequencies. The voltage gain as a

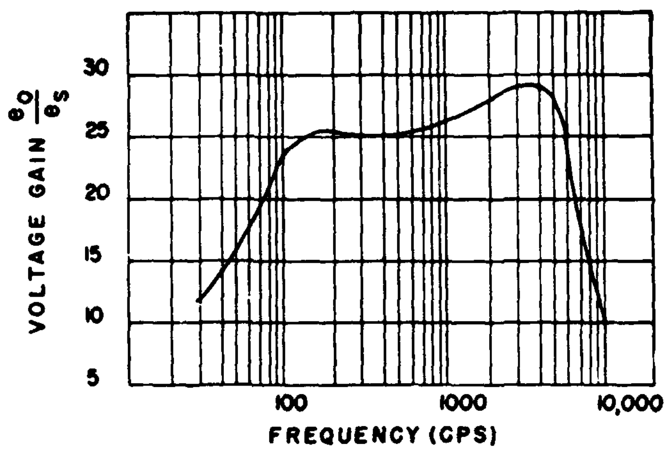


Figure 4-22.—Voltage gain vs frequency of a transformer-coupled voltage amplifier.

function of frequency throughout the range in question is shown in figure 4-22. The curve shows that the transformer-coupled voltage amplifier has a relatively high gain and uniform frequency response over the middle range of audio frequencies, but poor response for both low and high audio frequencies.

Like those of resistance-coupled amplifiers, complete equivalent circuits of transformer-coupled amplifiers are complex networks. An analysis of them can be consider-

ably simplified by considering one at a time the equivalent circuits for the low, middle, and high frequencies.

MIDDLE-FREQUENCY GAIN.—The primary of transformer T (fig. 4–21, A) is connected in the plate circuit of $V1$ and the secondary is connected between the grid and cathode of $V2$. An input signal, e_s, applied between the grid and cathode of $V1$, appears as an amplified plate signal, e_p, across the transformer primary.

At the middle frequencies the reactances of the transformer primary inductance, L_p, and secondary distributed capacitance, C_s (fig. 4–21, B), are sufficiently high to be considered as open circuits and there is no loss of signal voltage in the plate resistance of $V1$. Thus μe_s is equal to e_p. The secondary output voltage, e_i, applied to the input of $V2$, is equal to Ne_p or μNe_s, where N is the secondary-to-primary transformer turns ratio.

The simplified equivalent circuit applicable to the middle frequencies is shown in figure 4–23. In this figure the

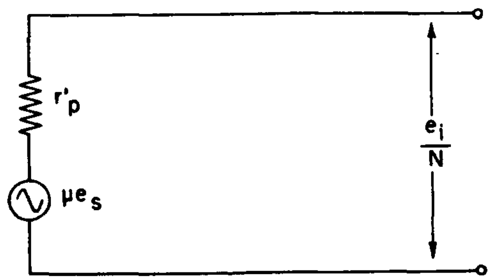


Figure 4–23.—Equivalent circuit of transformer-coupled voltage amplifier for mid-frequency operation.

primary inductance and secondary distributed capacitance have been omitted and resistor r'_p includes the plate resistance and the primary winding resistance in series. The transformer leakage reactances are negligible and the output voltage, $\frac{e_i}{N}$, is equal to μe_s since there is no drop through r'_p.

The middle-frequency gain is equal to $\frac{e_i}{e_s}$, or μN. The circuit applies to a class-A voltage amplifier in which no grid current flows during any part of the input cycle.

LOW-FREQUENCY LIMIT.—At the lower frequencies the shunting effect of the interelectrode and distributed capacitances is even less than it was at the middle frequencies (X_C varies inversely with the frequency), and is omitted in the equivalent circuit shown in figure 4–24. The reactance,

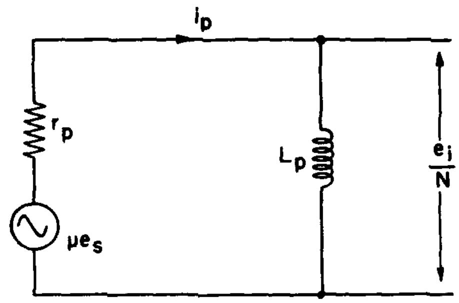


Figure 4–24.—Equivalent circuit of transformer-coupled voltage amplifier for low-frequency operation.

ωL_p, of the transformer primary is reduced at low frequencies.

The low frequency at which the gain falls to 70 percent of the middle frequency gain is the frequency at which ωL_p is equal to r'_p. Thus,

$$2\pi f L_p = r_p$$

and

$$f = \frac{r_p}{2\pi L_p}.$$

For example, an audio voltage amplifier triode has an r_p of 16,000 ohms and a μ of 8.5. The coupling transformer has a step-up voltage ratio of $\frac{3}{1}$ and a primary inductance of 40 henrys.

The low frequency limit is

$$f=\frac{16{,}000}{2\pi\times 40}=63.7 \text{ cps.}$$

The middle-frequency gain is μN, or $8.5\times 3=25.5$. Thus the low-frequency limit of 63.7 cps is the frequency at which the gain falls to 70 percent of 25.5, or about 17.7 (fig. 4–22).

The decrease in the reactance of the transformer primary inductance causes a falling off in gain at the lower frequencies. The falling off begins at higher frequencies when high-mu tubes having high r_p are used. The larger the transformer primary incremental inductance (inductance with partial d-c core saturation) the better is the low-frequency response.

HIGH-FREQUENCY LIMIT.—At the high-frequency end of the band the reactance of the primary inductance is high and is neglected in the equivalent circuit (fig. 4–25). The effect of

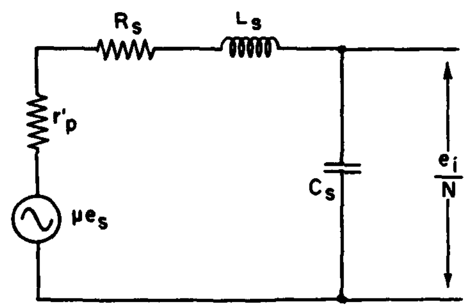


Figure 4–25.—Equivalent circuit of transformer-coupled voltage amplifier for high-frequency operation.

the shunting capacitance, C_s, is appreciable and the output voltage appears across it. The leakage inductance of the transformer acts in series with C_s, r_p, and the winding resistances.

These factors are included in the equivalent circuit for high-frequency operation which is similar to a series *L-C-R*

circuit having a low Q and operating in the vicinity of the series-resonant frequency. The plate resistance and the primary transformer winding resistance are included in r'_p and the secondary winding resistance expressed in terms of the primary side is equal to R_s. The total leakage inductance of both windings is equal to L_s expressed in terms of the primary. The shunting capacitance, C_s, is made up of the secondary-winding distributed capacitance and the input capacitance of the next tube, and is expressed in terms of the transformer primary.

Although the Q of this circuit is relatively low, the series-resonant effect is sufficiently pronounced to increase the gain at and near resonance unless precautions are taken to prevent this effect. Above resonance the gain falls off rapidly. Greatest uniformity of gain occurs in the high-frequency end of the band when the Q of the series-resonant circuit is approximately 0.85.

The high-frequency limit is in the vicinity of the series-resonant frequency, which is determined by the transformer leakage inductance and the effective shunting capacitance.

For example, the high-frequency end of the band of a transformer-coupled amplifier in which the leakage inductance of the transformer is 0.4 henry and the shunting capacitance is 1,500 micromicrofarads is

$$f_o = \frac{1}{2\pi\sqrt{L_s C_s}}$$

$$= \frac{1}{6.28\sqrt{0.4\times 1{,}500\times 10^{-12}}}$$

$$= 6{,}460 \text{ cps.}$$

Above this frequency the gain falls off rapidly because of the decrease in reactance of the shunting capacitance across which the output voltage is developed and also because of the increase in reactance of the leakage inductance in series with the output. Good high-frequency response is obtained by using a transformer having a small distributed capacitance.

This effect is obtained by using a low secondary-to-primary turns ratio which means a limited voltage gain, and a small transformer which means a poor low-frequency response. Thus transformer coupling is a compromise between high gain and good high- and low-frequency response.

Direct Coupling

In each of the coupling circuits that have been considered so far, the coupling device isolates the d-c voltage in the plate circuit of one tube from the grid circuits of the next tube; but they are designed to transfer the a-c component with minimum attenuation.

In a direct-coupled amplifier, on the other hand, the plate of one tube is connected directly to the grid of the next tube without going through a capacitor, a transformer, or any similar coupling device. This arrangement presents a problem of voltage distribution. Since the plate of a tube must have a positive voltage with respect to its cathode, and the grid of the next tube must have a negative voltage with respect to its cathode, it follows that the two cathodes cannot operate at the same potential. Proper voltage distribution is obtained by a voltage divider, as shown at *A*, *B*, *C*, *D*, and *E* in figure 4–26. In this amplifier the plate of *V*1

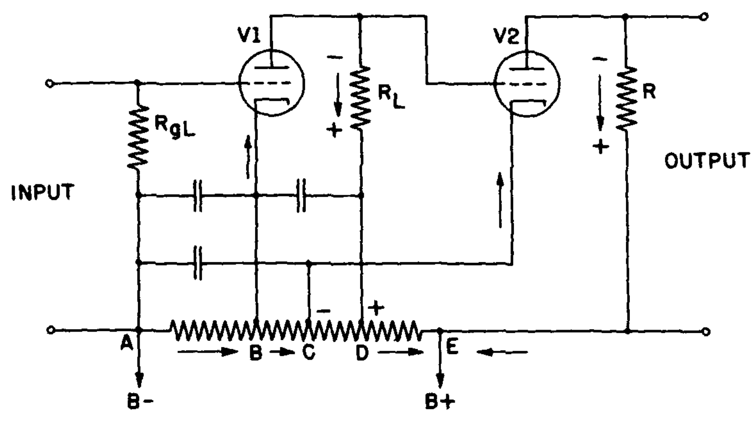


Figure 4-26.—Direct-coupled amplifier.

is connected directly to the grid of $V2$. The grid of $V1$ is returned to point A through R_{g1}. The cathode of $V1$ is returned to point B and the grid bias for $V1$ is developed by the voltage drop between points A and B of the voltage divider. The plate of $V1$ is connected through its plate load resistor, R_L, to point D on the divider. R_L also serves as the grid resistor for $V2$.

Since the plate current from $V1$ flows through R_L, a certain amount of the supply voltage appears across R_L. The amount of voltage developed across R_L must be allowed for in choosing point D on the divider. Point D is so located that approximately half of the available voltage is applied to the plate of $V1$. The plate of $V2$ is connected through a suitable output load, R, to point E, the most positive point on the divider. Since the voltage drop across R_L may place too high a negative bias on the grid of $V2$, it may be necessary to connect the cathode of $V2$ at point C, which is negative with respect to point D, in order to lower the bias on the grid of $V2$ (since the voltages across R_L and CD are in opposition). Point C, together with the value of R, determines the proper plate voltage for $V2$.

The entire circuit is a complex resistance network that must be adjusted carefully to obtain the proper plate and grid voltages for both tubes. If more than two stages are used in this type of amplifier, it is difficult to achieve stable operation. Any small changes in the voltages of the first tube will be amplified and will thus make it difficult to maintain proper bias on the final tube connected into the circuit. Because of the instability thus encountered, direct-coupled amplifiers are practically always limited to two stages. Furthermore, the power supply must be twice that required for one stage.

When the tube voltages are properly adjusted to give class-A operation, the circuit serves as a distortionless amplifier whose response is uniform over a wide frequency range. This type of amplifier is especially effective at the lower frequencies because the impedance of the coupling elements does not vary with the frequency. Thus a direct-

coupled amplifier may be used to amplify very low frequency variations in voltage. Also, because the response is practically instantaneous, this type of coupling is useful for amplifying pulse signals where all distortion caused by the coupling elements must be avoided.

QUIZ

1. Which component (a-c or d-c) of the voltages and currents in an amplifier circuit determines the portion of the tube characteristic at which operation occurs?
2. How may transformer coupling make the gain of a stage greater or less than mu?
3. What is the primary function of voltage amplifiers?
4. Express the formula for the gain of a voltage amplifier in terms of the output and input voltages.
5. Define power amplification.
6. Define plate efficiency.
7. Define power sensitivity.
8. During what part of the input cycle does plate current flow in a class-A amplifier?
9. What is the action of class-B audio amplifiers connected in push-pull?
10. During what part of the input cycle does plate current flow in a class-AB amplifier?
11. What is the principal use of class-C amplifiers?
12. Why must modified resistance coupling be used in video-frequency amplifiers?
13. What are the effects of grounding the grid and introducing the signal in the cathode circuit as shown in figure 4-8?
14. Give one advantage of the grounded-plate amplifier over the grounded-grid amplifier.
15. What ratio determines whether an amplifier is termed a wide-band or a narrow-band amplifier?

16. What type of distortion results when some frequency components of a signal are amplified more than others or when some frequencies are excluded?
17. What type of distortion is introduced when not all the frequency components of a signal are passed through an amplifier in the same amount of time?
18. What type of distortion occurs as a result of operating an electron-tube amplifier on the nonlinear portion of its characteristic curve?
19. What determines the ultimate sensitivity of an amplifier?
20. When resistance-coupled amplifiers are modified to *extend* the frequency range, what is the effect on the stage gain over the entire range?
21. In a resistance-coupled amplifier, what causes the gain to fall off at very low frequencies?
22. In the resistance-coupled amplifier shown in figure 4–17, B, what causes the gain to fall off at very high frequencies?
23. In a resistance-coupled amplifier, to what extent is the frequency response affected within the middle range of frequencies by the coupling capacitor, the output capacitance of one stage, the input capacitance of the next stage, and the distributed capacitance of the coupling network?
24. Name three advantages of using transformers as interstage coupling between voltage amplifiers.
25. What causes the reduced gain of a transformer-coupled amplifier at the lower audio frequencies?
26. Why does the gain of a transformer-coupled amplifier decrease at the higher audio frequencies?
27. In figure 4–26, what is the polarity of the cathode of *V2* with respect to the cathode of *V1*?
28. What are two of the good features of direct-coupled amplifiers?

CHAPTER 5

ELECTRON-TUBE AMPLIFIER CIRCUITS

DIRECT-CURRENT AMPLIFIERS

Operation

Direct-current amplifiers are used to amplify changes in direct current or voltage as well as very-low-frequency voltages. The simplest form of d-c amplifier consists of a single tube with a grid resistor across the input and a load connected in the plate circuit. The load may be an electromechanical device such as a meter, relay, or counter; or the output may be used to control the gain of an amplifier or the frequency of an oscillator, as in the automatic-frequency-control circuits of microwave receivers.

The d-c voltage change to be amplified is applied directly to the grid of the amplifier tube. Therefore, direct coupling (fig. 5–1, A) is required in the input circuit. A capacitor-input circuit (fig. 5–1, B) is also shown for comparison.

In the capacitor-input circuit of figure 5–1, B, graphs of the signal voltage, grid voltage, and plate current are shown above the circuit. The applied d-c voltage charges the capacitor, and momentarily the voltage drop across R_g equals the applied voltage; and this voltage then appears between the grid and cathode of the tube. However, when the capacitor is charged up to the value of the d-c input voltage; the current stops flowing through R_g and the grid reverts to its

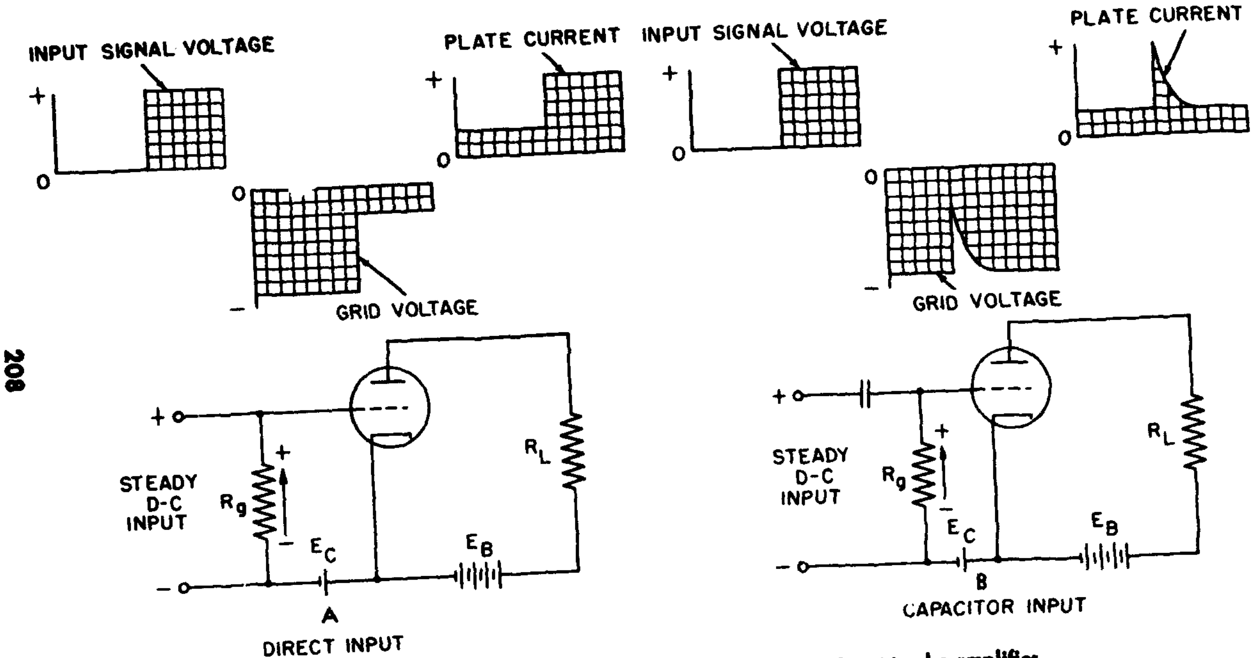


Figure 5–1.—Comparison of direct input and capacitor input to d-c amplifier.

original value, that of the bias voltage. Thus except for the original surge of plate current, which occurs when the capacitor is charging, there is no increase in voltage across R_L and hence no amplification.

In the direct-coupled input circuit of figure 5–1, A, the graphs of input signal, grid voltage, and plate current are shown above the circuit. The input signal is like that in figure 5–1, B, but here the similarity ends. With no input signal the negative bias voltage is present on the grid of the tube and a steady value of plate current flows. This action causes a fixed voltage drop across R_L. When a direct voltage of the polarity indicated is applied across the input terminals, there is no blocking action by a capacitor as in the previous case. Instead, the applied signal continues as a steady voltage drop across R_g canceling a portion of the negative bias. The net bias then drops to the new value indicated in the grid-voltage graph. This reduction in grid bias causes a greater current flow in the plate circuit, and thus a greater drop appears across R_L. Thus, the increase in plate current is sustained as long as the input signal voltage exists at the corresponding level that caused the plate current to increase.

Use

One of the most important applications of a d-c amplifier is its use as a d-c electron-tube voltmeter. A typical circuit, equivalent to the one in figure 5–1, A, is shown in figure 5–2.

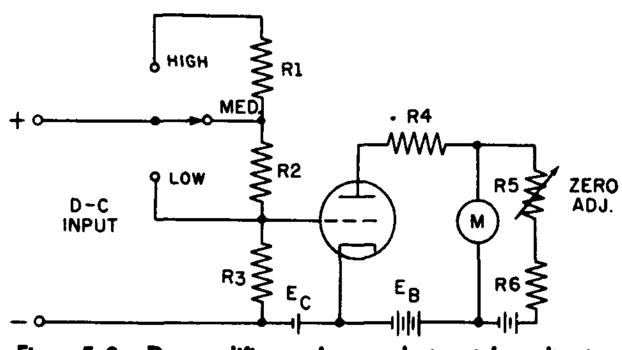


Figure 5–2.—D-c amplifier used as an electron-tube voltmeter.

The d-c voltage to be measured is applied via the range switch to one of the three taps on the voltage divider which is made up of $R1$, $R2$, and $R3$. The purpose of resistor $R4$ is to prevent damage to the tube in the event that too high a voltage is applied to the input. In the plate circuit an additional d-c voltage and two resistors, $R5$ and $R6$, are used to balance the meter current to zero when no voltage is applied to the input. One of the resistors, $R5$, is variable and permits zero adjustment of the meter under no-signal conditions by bucking out the voltage applied across the meter by E_B.

If a d-c voltage is then applied to the input, current flows through the meter. This current is proportional to the applied voltage and its value can be read directly on the calibrated scales (one for each range) of the meter. Various modifications permit this basic circuit to be utilized for measuring other quantities, or even to be incorporated in a multitester.

Another important application of the d-c amplifier is shown in figure 5-3. Here the amplifier is employed in two of the legs of a bridge circuit that is used to show the exact point of balance between two d-c voltages. If the tubes are

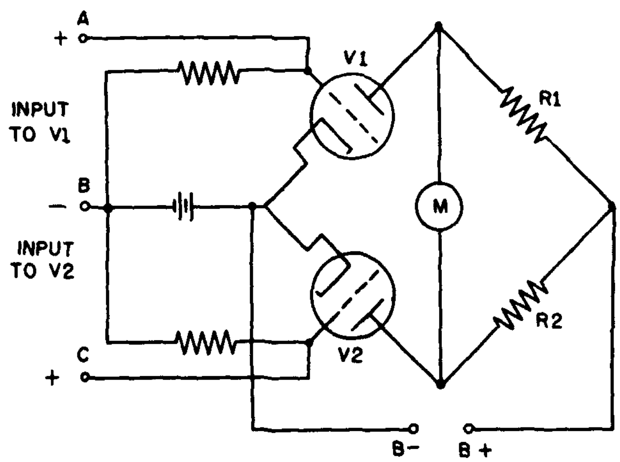


Figure 5-3.—Balanced d-c amplifier.

properly matched and if there is no input signal between terminals A and B or B and C, no current flow is indicated through the meter since the IR drops across $R1$ and $R2$ are equal. When unequal signals of the polarity indicated are applied between A and B and between B and C, the grid-to-cathode voltages of the two tubes are unbalanced. This action unbalances the plate currents and the IR drop across $R1$ and $R2$. Thus the bridge becomes unbalanced and the two d-c voltages may be compared if they are applied simultaneously with the polarities indicated.

The amount of current indicated by the meter is proportional to the difference between the two applied voltages. Since the meter has its zero position in the middle of the scale, the direction that the needle swings indicates which voltage is greater. This type of amplifier circuit has many radio and radar applications.

A specific application of a d-c amplifier is its utilization in automatic-frequency-control (a-f-c) circuits in microwave receivers. Briefly, in such receivers the center frequency of the i-f signal must be constant. Any slight variations in the transmitter frequency or the local oscillator frequency are balanced out by varying the potential on the proper electrode of the local oscillator. The intermediate frequency is equal to the difference between the transmitter frequency and the local oscillator frequency and will vary when either the transmitter frequency or the local oscillator frequency varies. A discriminator circuit detects these variations in the i-f signal frequency and feeds a d-c voltage, which is proportional to the variations, to a d-c amplifier. The amplifier output is fed to the proper oscillator electrode in such a way as to correct for the variation. There are many more important applications of d-c amplifiers in communications, radar, and industry.

FEEDBACK AMPLIFIERS

As the term implies, a voltage feedback amplifier transfers a voltage from the output of the amplifier back to its input. If the signal is fed back in phase with the input signal it is

called POSITIVE, DIRECT, or REGENERATIVE feedback because it adds to the voltage of the input. If the signal fed back to the input is 180° out of phase with the applied signal it is called NEGATIVE, INVERSE, or DEGENERATIVE feedback because it subtracts from the input voltage.

Principle of the Feedback Amplifier

The principle of the feedback amplifier may be understood in part from a consideration of figure 5–4. A signal voltage,

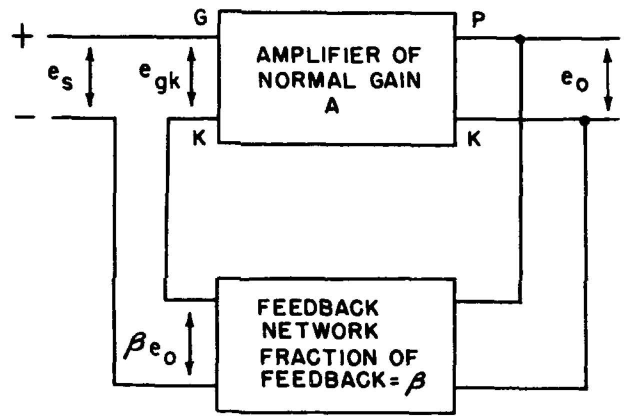


Figure 5–4.—Feedback employing series injection.

e_s, is applied to the input terminals, as shown in the figure. Let a portion, βe_o, of the output voltage, e_o, be fed back in series with e_s in such a way that the signal, e_{gk}, appearing between the grid and cathode is of the form

$$e_{gk}=e_s+\beta e_o, \tag{5–1}$$

where β is a fractional part of e_o.

Since the normal gain, A, of the amplifier is defined as

$$A=\frac{e_o}{e_{gk}},$$

then

$$e_o = Ae_{gk}. \tag{5-2}$$

Substituting the value of e_{gk} from equation (5-1) in equation (5-2) gives

$$e_o = A(e_s + \beta e_o),$$

and

$$e_o = \frac{Ae_s}{1-\beta A}.$$

Negative Feedback

If βA is greater than 1, the quantity $1-\beta A$ is negative and the amplification is less than it would be without feedback. Thus the amplification is said to be negative or degenerative. Generally the feedback factor, βA, is so much larger than 1 that the resultant amplification A_r for all practical purposes may be expressed as

$$A_r = -\frac{1}{\beta}. \tag{5-3}$$

Advantages.—Negative feedback may be used to reduce the nonlinear distortion—that is, to make the output waveform more nearly similar to the input waveform by reducing nonlinearities that are introduced within the amplifier tube itself. This use may be understood by the following considerations.

The input signal applied to the grid of an electron-tube amplifier is amplified by an amount determined by the μ of the tube, but any nonlinearities introduced within the tube are not amplified. If a portion, βA, of the output is fed back 180° out of phase with the input, the distortion component of this feedback voltage will be amplified along with the input signal. The amplified distortion component will tend to cancel the distortion component introduced within the tube, and the output may be practically free of nonlinear distortion. It is necessary that the distortion occur in the plate circuit of the stage across which negative feedback is to be applied, in order to separate the distortion from the desired signal.

However, the over-all gain of the desired signal will also be reduced; but this reduction may be compensated for by increasing the number of stages. Distortion caused by the flow of grid current cannot be corrected by negative feedback because this distortion occurs at the source and not within the amplifier tube.

The resulting gain, A_r (with feedback considered), is

$$A_r = \frac{e_o}{e_s};$$

therefore,

$$A_r = \frac{A}{1-\beta A}.$$

The resultant amplifier gain is expressed in terms of the gain without feedback, A, and the fraction of the output, β, fed back to the input. A_r, A, and β may be complex quantities.

Positive Feedback

If the quantity $1-\beta A$ is less than 1, the gain of the amplifier is increased over what it would be without feedback, and the amplifier is said to be positive or regenerative. Under these conditions the response curve is sharpened and the gain is increased, but the frequency range of uniform response is reduced. Thus positive feedback affords both an increase in gain and an increase in selectivity.

The increase in gain, however, is accompanied by an exaggeration of any undesirable distortion or noise that was introduced within the amplifier itself. For this reason positive feedback is not used if a distortionless output is required. If the feedback factor, βA, is increased until it is equal to 1, the quantity $1-\beta A$ reduces to 0; and the resultant gain theoretically becomes infinite, or at least large enough to sustain oscillations. Under this condition no input voltage is required to obtain an output voltage and the amplifier becomes an oscillator. This aspect of feedback is discussed in chapter 7 on oscillators.

Noise introduced within an amplifier may be reduced by negative feedback in the same manner that nonlinear distortion was reduced; and the same limitations apply—that is, for feedback to be effective, the noise to be canceled out must be that which is generated in a tube around which the feedback is applied. Thus, thermal agitation, induced hum, microphonics, and shot effects introduced in the early stages of a receiver cannot be reduced by negative feedback unless the feedback is applied at those stages.

Negative feedback in those stages would not be practical because the amplification, particularly at the high frequencies encountered in most radar receivers, is low and negative feedback would reduce it even more. Additional stages, each with its own circuit noises, would have to be added to make up for the reduced gain. Negative feedback is very effective, however, in reducing noises, particularly hum, introduced in the high-level (high-power) stages of an amplifier. If that part of the output voltage fed back to the input is obtained by means of a resistance network the resultant amplification is essentially independent of frequency.

When it is desired to have the amplification vary in some specific manner with respect to frequency, the negative feedback network through which β is obtained may be designed to attenuate those frequencies that are desired in the output of the amplifier. For example, if the high frequencies are to be amplified more than the low frequencies, the high frequencies must be attenuated in the feedback network more than the low frequencies. Only the low frequencies will be fed back to the grid in phase opposition to the input signal and will therefore be reduced in the output.

The over-all gain of a negative feedback amplifier can be made substantially independent of the magnitude of the load impedance provided the load impedance does not interfere with the feedback signal. This may be understood if it is assumed, for example, that as the effective load resistance is reduced, the a-c component of the plate voltage tends to decrease. Accordingly there is less negative feedback and the amplitude of the grid signal is increased. Thus, the

increased grid signal offsets the tendency of the output voltage to drop and the over-all gain is approximately constant. On the other hand, if the effective resistance of the load is increased, the a-c component of plate voltage will tend to increase and the negative feedback will increase. Thus the amplitude of the grid signal is decreased and the tendency for the output voltage to rise is checked. Again, the over-all gain is held approximately constant.

Since the gain is proportional only to the feedback factor, the gain is independent also of such factors as variations in supply voltages or aging of tubes.

In negative-feedback amplifiers nonlinear distortion, noise originating within the tube, and frequency distortion may be reduced. In other words, amplitude and phase characteristics can be corrected by means of negative feedback. Likewise, the effects of variations in load and plate voltage supply, as well as the effects of tube aging, may be effectively counteracted. The price paid for these advantages is a reduction in gain—that is, an increase in the number of stages.

Methods of obtaining negative feedback.—In a practical amplifier negative feedback may be obtained in a number of ways and it may involve one or two stages and in rare instances more than two stages. Also, it may employ voltage feedback, current feedback, or a combination of voltage and current (compound) feedback.

Voltage feedback is obtained by means of a voltage divider network, *C1R1R2*, as shown in figure 5–5. In this circuit a part of the output voltage appearing across the primary of the output transformer, *T2*, is fed back through the coupling network, *C1R1*, in series with the secondary of the input transformer, *T1*, as the voltage drop across *R2*.

To analyze the action, assume that the input signal causes the grid voltage to swing in a positive direction. The plate voltage falls as plate current increases. During this time capacitor *C1* discharges through *R1* and *R2*. The voltage drop across *R2* has a polarity that makes the top end of *R2* negative with respect to ground. Thus the voltage across

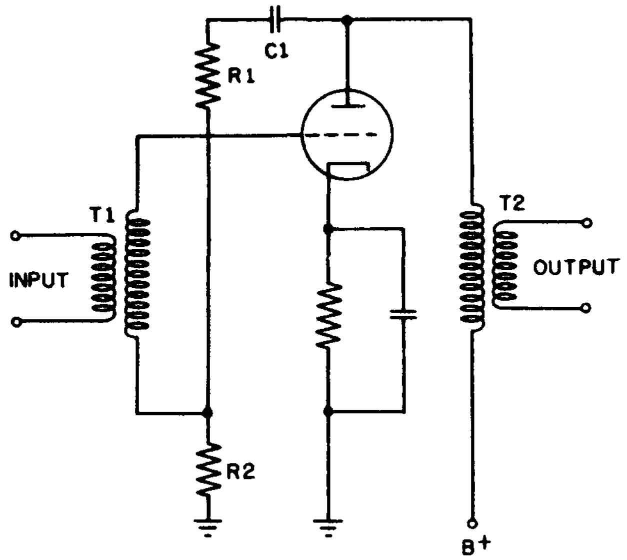


Figure 5–5.—Degenerative amplifier employing voltage feedback.

*R*2 is in phase opposition with respect to the secondary voltage of *T*1. The grid-cathode signal is therefore equal to the difference in the secondary voltage of *T*1 and the feedback voltage across *R*2.

Another method of obtaining negative feedback is illustrated in figure 5–6. This method employs current feedback. Here the cathode resistor bypass capacitor has been omitted. The degenerative action may be analyzed as follows: Assume that the input signal swings the grid voltage in a positive direction. The increase in plate current causes an increase in the voltage drop across R_k. Since R_k is not bypassed, plate circuit signal currents flowing through R_k will add to the bias produced by the no-signal component. The grid-to-cathode voltage on the positive half cycle is equal to the difference in the input and the drop across R_k. Thus the magnitude of the grid voltage swing in a positive direction is not as great as it would be without feedback because the drop across R_k is increased.

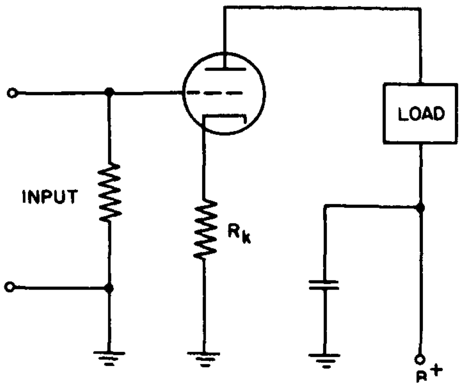


Figure 5–6.—Degenerative amplifier employing current feedback.

Similarly on the negative half cycle the input signal swings the grid voltage in a negative direction and plate current decreases. The decrease in current through R_k causes a decrease in the voltage across R_k. During this half cycle the grid-to-cathode voltage is equal to the sum of the input voltage and the drop across R_k. Thus the magnitude of the negative swing of grid voltage is less than it would be without feedback because the drop across R_k is less.

The fact that an output voltage, opposite in phase to the input voltage, may be developed across an unbypassed cathode resistor is used in designing cathode followers and phase inverters. These circuits are considered later in this chapter.

If proper phase relations are established, negative feedback involving more than one stage may be employed. Figure 5–7 shows a 2-stage negative feedback amplifier employing voltage feedback. In this case special attention must be paid to the phase relations throughout the circuit.

Assume that at a given instant the input voltage is such as to make the grid of $V1$ less negative. Plate current then increases in $V1$ and the plate voltage decreases, causing the grid of $V2$ to become more negative. At the same time the plate of $V2$ becomes more positive because of the reduction in

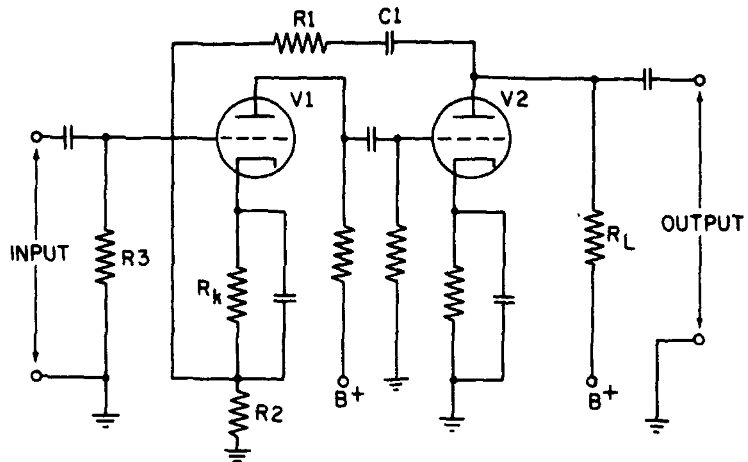


Figure 5-7.—Degenerative 2-stage amplifier employing voltage feedback.

plate current. This increase in potential causes an increase in the charge of $C1$. The charging current flows from ground up through $R2$ and $R1$ to the left plate of $C1$, making the top end of $R2$ more positive with respect to ground. The increase in voltage across $R2$ acts in series with the input and the bias across R_k to reduce the magnitude of the positive-going signal impressed on the grid. In short, the grid input signal is reduced by the amount of the feedback voltage because these two voltages are 180° out of phase.

Various combinations of voltage and current feedback circuits may be employed to satisfy specific requirements. Thus, a compound feedback employing both current and voltage feedback may be employed in a single stage, or current feedback may be used in one stage of a 2-stage amplifier section, and in addition voltage feedback may be employed between the stages.

TUNED AMPLIFIERS

A part of the coupling circuit of a tuned amplifier is a parallel resonant circuit. Such circuits are used because

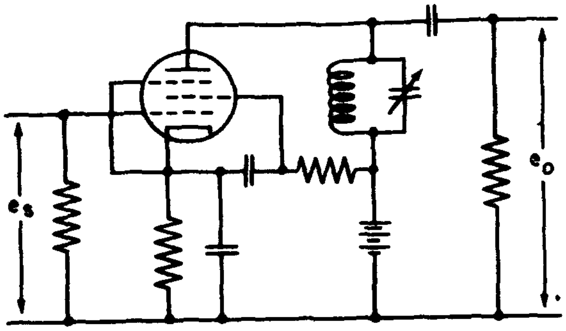
they offer high impedance at the desired frequency and low impedance at other frequencies, thus permitting the amplification of a relatively narrow band of frequencies. In addition, the limitations imposed on untuned amplifiers by interelectrode and distributed capacitances are used to advantage because these capacitances become a part of the tuned circuit. These amplifiers may be single- or double-tuned depending on whether the plate circuit only or both the plate and the grid circuits contain parallel-tuned circuits.

The three basic tuned-amplifier circuits are shown in figure 5–8. Each circuit has specific applications. For example: The single-tuned resistance-coupled amplifier (fig. 5–8, A) might be used with various modifications to deliver energy from a class-C power amplifier; the single-tuned transformer-coupled amplifier circuit (fig. 5–8, B) might be used as a class-A r-f voltage amplifier (preselector) ahead of the first detector in a superheterodyne receiver; and the double-tuned transformer-coupled amplifier circuit (fig. 5–8, C) might be used as an intermediate frequency stage of amplification between the first and second detectors in the superheterodyne receiver.

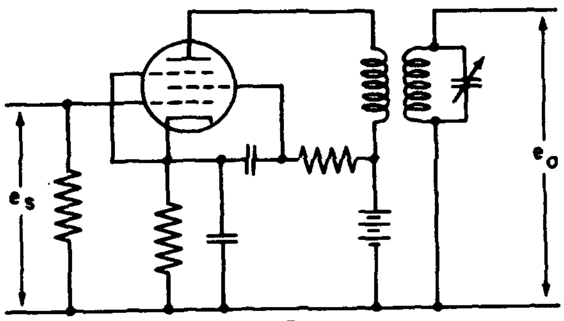
A better understanding of how these circuits operate may be had if the equivalent circuit of each basic amplifier is presented along with a brief analysis of the circuit.

Analysis of Single-Tuned R-C Coupling

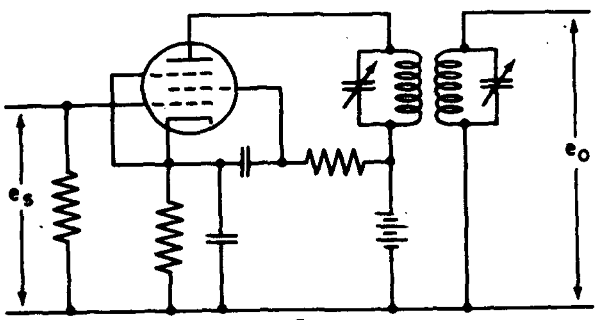
The simplified equivalent circuit of figure 5–8, A, is shown in figure 5–9. In figure 5–9, capacitor C includes the capacitance of the tank tuning capacitor, stray capacitances, and the output and input tube capacitances. The capacitance of the series coupling capacitor is large and its impedance is negligible. Thus it has been omitted from the equivalent circuit. The tank coil, L, resonates with C at the desired frequency. The plate resistance, r_p, of the amplifier stage and the grid-leak resistance, R_g, of the next



A
SINGLE-TUNED RESISTANCE COUPLED



B
SINGLE-TUNED TRANSFORMER COUPLED



C
DOUBLE-TUNED TRANSFORMER COUPLED

Figure 5-8.—Basic tuned-amplifier circuits.

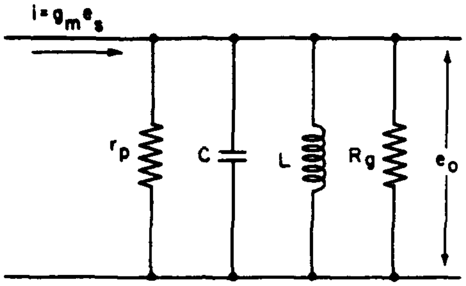


Figure 5–9.—Equivalent circuit of a single-tuned amplifier—constant-current generator form.

stage are in parallel with the tank. The amplification or voltage gain (A) is

$$A=\frac{e_o}{e_s}=\frac{iZ_L}{e_s}=\frac{g_m e_s Z_L}{e_s}=g_m Z_L,$$

where Z_L is the combined impedance of the parallel circuit.

At resonance, the impedance of the parallel resonant tank, CL, is large and resistive and is equal to Q times the reactance of either C or L. Therefore,

$$R_o=X_L Q=\omega LQ=2\pi f_o LQ,$$

where Q is the effective value of the tank circuit alone (r_p and R_g omitted) and f_o is the resonant frequency.

The total impedance of the circuit, including the impedance of the resonant tank, CL, in parallel with r_p and R_g is

$$Z_L=\frac{1}{\frac{1}{r_p}+\frac{1}{\omega LQ}+\frac{1}{R_g}}$$

$$=\frac{\omega LQ}{1+\frac{\omega LQ}{r_p}+\frac{\omega LQ}{R_g}}.$$

The amplification at resonance becomes

$$A = \frac{g_m \omega L Q}{1 + \frac{\omega L Q}{r_p} + \frac{\omega L Q}{R_g}}.$$

Normally, for the type of tube generally used, R_g is high with respect to the impedance, ωLQ, of the resonant tank, and for pentodes, r_p is very high. The approximate amplification may be expressed as

$$A \cong g_m \omega L Q.$$

Analysis of Single-Tuned Transformer Coupling

The derivation of an expression for the voltage gain of the single-tuned transformer-coupled amplifier may be more easily understood if the following considerations are stated first: The voltage induced in the secondary, $L2$, by the primary current, i_p, is equal to $\omega M i_p$, where M is the mutual inductance between the primary and the secondary. This induced voltage is multiplied by a factor, Q, to obtain the output voltage across C because of the resonant voltage rise in the tuned circuit. (See fig. 5–10).
The output is now established as

$$e_o = Q \omega M i_p.$$

Because r_p is large with respect to ωL_1 and the effect of the presence of the tuned secondary on the primary is slight (low coupling), the primary current is dependent upon μe_s and r_p. The expression for primary current then becomes

$$i_p = \frac{\mu e_s}{r_p}$$

and, therefore,

$$e_o = \frac{Q \omega M \mu e_s}{r_p}.$$

Since $\frac{\mu}{r_p}$ is equal to g_m,

$$e_o = Q \omega M g_m e_s$$

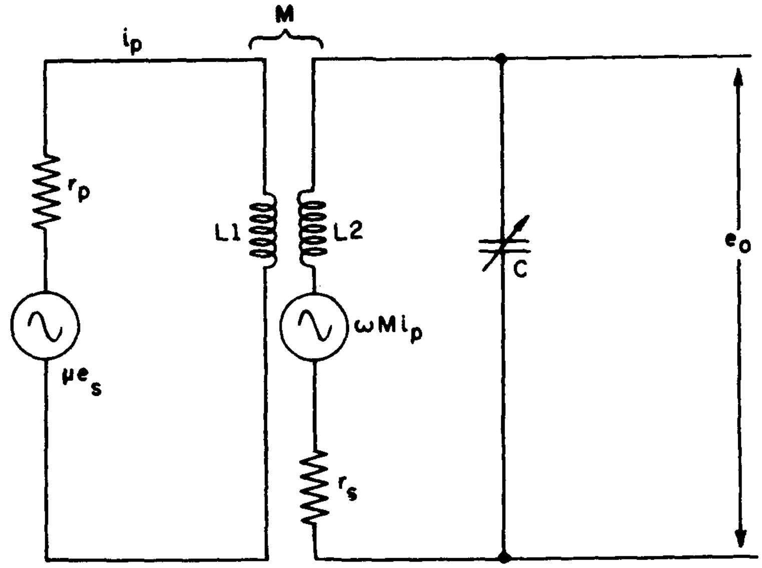


Figure 5-10.—Equivalent circuit of a single-tuned transformer-coupled amplifier—constant-voltage generator form.

and

$$\text{gain} = \frac{e_o}{e_s} = \frac{Q\omega M g_m e_s}{e_s} = Q\omega M g_m.$$

Thus, a high-gain amplifier has a low-loss tuned circuit and employs a tube having a high mutual conductance.

Analysis of Double-Tuned Transformer Coupling

As indicated in figure 5-11, the double-tuned transformer-coupled amplifier has a band-pass characteristic which depends in part on the degree of coupling and in part on the circuit Q's.

Under proper operating conditions essentially uniform amplification of a relatively narrow band of frequencies may be achieved and amplification of frequencies outside this

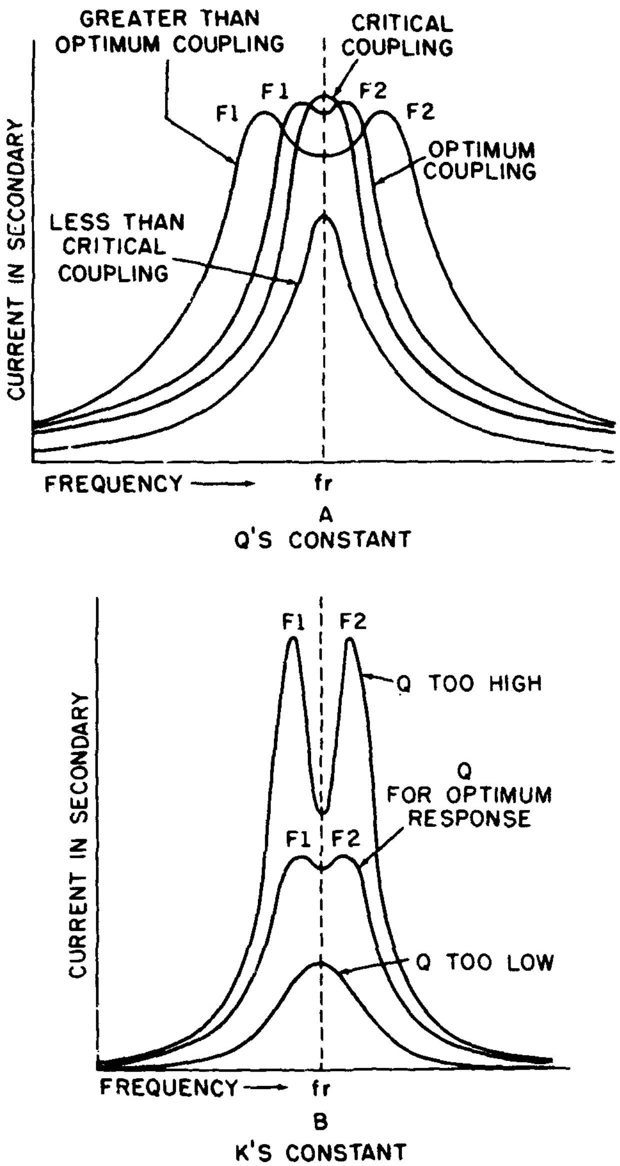


Figure 5-11.—Response curves for a double-tuned transformer-coupled amplifier.

band may be sharply reduced. These characteristics make this type of coupling highly desirable in intermediate amplifiers in both radio and television receivers.

Since the slope of the response curve is not perfectly vertical, the circuit cannot completely discriminate against frequencies just outside the desired channel without also attenuating to some extent the frequencies at the upper and lower limits of the pass band. However, double-tuned amplifiers approach an ideal band-pass characteristic much more closely than single-tuned amplifiers, which have rounded response curves.

To aid in the derivation of the gain formula for the double-tuned transformer-coupled circuit, the concept of coupled impedance is utilized. The effect of the presence of the tuned secondary on the primary is the same as if an impedance, $\frac{(\omega M)^2}{Z_s}$, had been added in series with the primary. This quantity is known as COUPLED IMPEDANCE.

The coupled impedance, Z_C (fig. 5–12, A), acting in series with the primary and caused by the presence of the secondary, is equal to the ratio of the voltage induced in the primary by the secondary, to the primary current. Thus,

$$Z_C=\frac{e_{ind}}{i_p}; \tag{5–4}$$

and by definition,

$$e_{ind}=\omega M i_s. \tag{5–5}$$

Substituting e_{ind} from equation (5–5) in equation (5–4),

$$Z_C=\frac{\omega M i_s}{i_p}. \tag{5–6}$$

By Ohm's law,

$$i_s=\frac{e_s}{Z_s}. \tag{5–7}$$

If the value of i_s from equation (5–7) is substituted in equation (5–6),

$$Z_C=\frac{\omega M\frac{e_s}{Z_s}}{i_p}. \qquad (5\text{–}8)$$

By definition,

$$e_s=\omega M i_p; \qquad (5\text{–}9)$$

and if the value of e_s from equation (5–9) is substituted in equation (5–8),

$$Z_C=\frac{\omega M\frac{\omega M i_p}{Z_s}}{i_p}=\frac{(\omega M)^2}{Z_s}.$$

It is helpful in analyzing the circuits of double-tuned transformer coupling to deal with the secondary as if it were coupled in series with the primary.

An expression for the gain of a double-tuned transformer-coupled pentode amplifier (fig. 5–12, B) may be derived by the use of the equivalent circuit shown in figure 5–12, C.

The following symbols are used in this derivation:

e_g-------- signal voltage applied to pentode grid
C_1-------- capacitance of capacitor $C1$
C_2-------- capacitance of capacitor $C2$
L_1-------- inductance of primary $L1$
L_2-------- inductance of secondary $L2$
e_o-------- output voltage across capacitor $C2$ due to resonance
i_1-------- current in primary
i_2-------- current in secondary
M-------- mutual inductance of $L1$ and $L2$
ω-------- $2\pi f$
e_s-------- voltage induced in secondary by transformer action
e_p-------- equivalent signal voltage acting in equivalent series circuit
Z_s-------- series impedance of secondary
R_2-------- series impedance of secondary resonance
Z_p-------- series impedance of primary before coupling secondary to it
Z_c-------- impedance coupled into the primary circuit by secondary
r_p-------- pentode plate resistance
R_{eq}-------- equivalent series resistance of pentode plate resistance
k-------- coefficient of coupling
Q_p-------- effective Q of primary circuit
Q_s-------- effective Q of secondary circuit
X_{C1}-------- capacitive reactance of $C1$
X_{C2}-------- capacitive reactance of $C2$

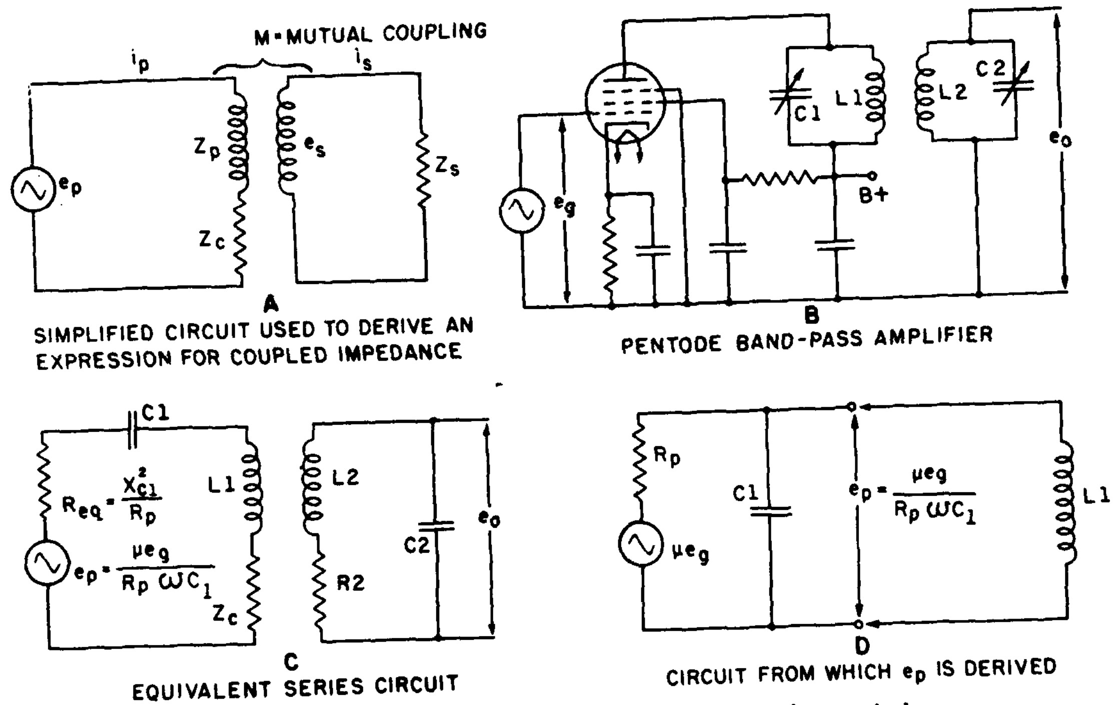


Figure 5-12.—Circuits used for deriving an expression for voltage gain in a pentode i-f band-pass circuit.

The current in the secondary (fig. 5–12, C) is

$$i_2=\frac{e_s}{Z_s}=\frac{\omega M i_1}{Z_s}, \tag{5–10}$$

since the voltage induced in the secondary of a transformer by the primary current, i_1, has a magnitude of $\omega M i_1$.

The primary current is

$$i_1=\frac{e_p}{Z_p+Z_c}, \tag{5–11}$$

where e_p is the equivalent voltage acting in the equivalent series circuit. This voltage is derived as the voltage appearing across $C1$ (fig. 5–12, D) when $L1$ is removed from the circuit. Thus,

$$e_p=iX_{C1} \tag{5–12}$$

and, since $r_p \gg X_{C1}$,

$$i=\frac{\mu e_g}{r_p}, \tag{5–13}$$

where i is the current in $C1$ when $L1$ is removed from the circuit. Substituting the value of i from equation (5–13) in equation (5–12),

$$e_p=\frac{\mu e_g}{r_p \omega C_1}.$$

Substituting the value of i_1 from equation (5–11) into equation (5–10) and simplifying,

$$i_2=\frac{\omega M \dfrac{e_p}{Z_p+Z_c}}{Z_s}=\frac{\omega M e_p}{Z_p Z_s+Z_c Z_s}. \tag{5–14}$$

The impedance, Z_c, is

$$Z_c=\frac{(\omega M)^2}{Z_s}. \tag{5–15}$$

Substituting the value of coupled impedance from equation (5–15) into equation (5–14),

$$i_2=\frac{\omega M e_p}{Z_p Z_s+(\omega M)^2}. \tag{5–16}$$

At resonance, Z_p and Z_s are both resistive. Z_p is equal to $\frac{\omega L_1}{Q_p}$. Z_s is equal to $\frac{\omega L_2}{Q_s}$. M is equal to $k\sqrt{L_1L_2}$. Substituting these values in equation (5–16),

$$i_2=\frac{\omega k\sqrt{L_1L_2}\,e_p}{\left(\frac{\omega L_1}{Q_p}\right)\left(\frac{\omega L_2}{Q_s}\right)+\omega^2k^2L_1L_2}. \tag{5–17}$$

Substituting g_m for $\frac{\mu}{r_p}$ in equation (5–13) and the value of i from equation (5–13) into equation (5–12), the equivalent signal voltage acting in the equivalent series primary circuit is

$$e_p=g_m e_g X_{C1}. \tag{5–18}$$

The output voltage is

$$e_o=i_2X_{C2}. \tag{5–19}$$

Substituting the value of e_p from equation (5–18) in (5–17) and the value of i_2 from equation (5–17) in (5–19), the output voltage becomes

$$e_o=\frac{\omega k\sqrt{L_1L_2}\,g_m e_g X_{C1}X_{C2}}{\left(\frac{\omega L_1}{Q_p}\right)\left(\frac{\omega L_2}{Q_s}\right)+\omega^2k^2L_1L_2} \tag{5–20}$$

At resonance $X_{C1}=X_{L1}$ and $X_{C2}=X_{L2}$. Simplifying equation (5–20), the output voltage becomes

$$e_o=\frac{\omega k\sqrt{L_1L_2}\,g_m e_g X_{C1}X_{C2}}{X_{L1}X_{L2}\left(\frac{1}{Q_pQ_s}+k^2\right)}$$

$$= \frac{\omega k \sqrt{L_1 L_2}\, g_m e_g}{\frac{1}{Q_p Q_s} + k^2}. \tag{5-21}$$

Therefore, at resonance, the voltage gain, (A) is

$$\text{gain} = \frac{e_o}{e_g} = \frac{g_m \omega k \sqrt{L_1 L_2}}{\frac{1}{Q_p Q_s} + k^2} \tag{5-22}$$

From equation (5–22) it is apparent that the gain at resonance varies directly with g_m, and therefore the value of g_m should be high to keep the gain satisfactorily high. The gain also depends upon the Q of the circuits. In general, adding turns to a coil increases the inductive reactance in proportion to the square of the turns and the effective resistance more nearly in proportion to the first power of the turns so that increased inductance is equivalent to an increase in Q. For a given resonant frequency the L-C product is fixed so that an increase in L requires a decrease in C and the L-C ratio is increased. Thus an increase in the L-C ratio is equivalent to an increase in Q and amplifier gain.

To illustrate the order of magnitude of the various quantities involved in equation (5–22), the gain of an intermediate frequency amplifier is calculated by substituting the given values into the equation.

The constants of the i-f amplifier circuit are as follows:

$C_1 = C_2 = 94 \mu\mu f$
$L_1 = L_2 = 4{,}000 \mu h$
$g_m = 1{,}600$ micromhos
$\mu = 1{,}200$
$r_p = 800{,}000$ ohms
Coil Q's $= 50$
$k = 0.03$
$M = 120 \times 10^{-6}$ henry
$f_o = 260$ kc
$e_g = 0.096$ mv

$$X_{L1}=\omega L_1=6.28\times 260\times 10^3\times 4{,}000\times 10^{-6}=6{,}530 \text{ ohms}$$

$$R_{eq}=\frac{X^2_{L1}}{r_p}=\frac{(6{,}530)^2}{800{,}000}=53.2 \text{ ohms}$$

$$\text{Coil resistance}=\frac{X_L}{Q}=\frac{6{,}530}{50}=130.6 \text{ ohms}$$

$$Q_p=\frac{6{,}530}{130.6+53.2}=35.6 \text{ (taking into account the effect of the pentode)}$$

The voltage gain at resonance is (equation 5–22)

$$\text{gain}=\frac{1{,}600\times 10^{-6}\times 6.28\times 260\times 10^3\times \quad 0.03\sqrt{4\times 10^{-3}\times 4\times 10^{-3}}}{\frac{1}{35.6\times 50}+(0.003)^2}=214.$$

Thus a pentode amplifier having an amplification factor of 1,200 provides a voltage gain of approximately 214 at an intermediate frequency of 260 kc.

A better understanding of the gain at resonance, as well as the response throughout the pass band of a double-tuned transformer-coupled stage, may be gained from a further consideration of the curves of secondary current versus frequency, as shown in figure 5–11.

When the coefficient of coupling in figure 5–11, A, is low, the response is sharply peaked at the resonant frequency and the pass band is very narrow. As the coupling is increased to the critical value, maximum current flows in the secondary, and the output voltage across the secondary is also at its maximum. At this point (critical coupling)

$$k=\frac{1}{\sqrt{Q_pQ_s}};$$

and, if the Q's are equal,

$$k=\frac{1}{Q}.$$

The pass band is still relatively narrow and would attenuate

the side band frequencies farthest removed from the resonant frequency.

If the coupling is further increased until the optimum value is reached the gain is still relatively high; but the pass band has been increased and the response is essentially uniform. At this point (optimum coupling)

$$k=\frac{1.75}{\sqrt{Q_pQ_s}};$$

and, if the Q's are equal,

$$k=\frac{1.75}{Q}$$

As the coupling is again increased the humps at F_1 and F_2 are well defined and the gain at resonance is considerably reduced. Although the pass band is now much wider, the gain throughout the band is not sufficiently uniform.

The two humps in the curve are due to the reactance that is coupled into the primary on each side of resonance as the coupling is increased. Below resonance this reactance is inductive, and above resonance it is capacitive. For the same frequency the coupled reactance has the opposite sign to that of the primary and the impedance of the primary is therefore reduced. Accordingly there is an increase in primary current at frequencies slightly off resonance, and a corresponding increase in secondary induced voltage and current at these frequencies.

The width of the pass band may be as important as the response within the band. The approximate width is

$$\text{width of pass band}=kf_o,$$

where f_o is the resonant frequency which is the center of the pass band.

The frequencies at the two humps, F_1 and F_2, which define the practical lower and upper limits of the pass band are determined by the following equations:

$$F_1 = \frac{f_o}{\sqrt{1+k}};$$

$$F_2 = \frac{f_o}{\sqrt{1-k}}.$$

Figure 5–11, B, shows the effects of varying the Q while maintaining a constant coefficient of coupling. Actually, the desired response curve could be achieved by the proper manipulation of both k and Q because they are interrelated.

From the foregoing equations it is seen that in order for the pass band to be wide, k must be large and the circuit Q's small. However, the proper relation between k and the Q's is essential if both the desired bandwidth and the desired response within the bandwidth are to be maintained.

To increase k, the coils are brought closer together. To lower the Q's, the coils are shunted with suitable resistors. The lower the shunting resistance, the lower will be the circuit Q.

Modifications for High-Frequency Operation

When the frequency to be amplified is very high, conventional r-f and i-f amplifiers require special circuit modifications. The circuit of a high-frequency tuned amplifier used at radar frequencies and requiring such modification is shown in figure 5–13. Tuning is accomplished by means of a small coil which resonates with its own distributed capacitance and the interelectrode capacitance of the tube. It is tuned by either a brass slug which acts as a short-circuited turn or by a powdered iron core. Both methods vary the effective inductance of the coil and thus the resonant frequency of the amplifier stage.

Other features of this amplifier may be summarized as follows: The interelectrode capacitance of the tube must be low so that the coil will be large enough to be tuned conveniently. Since a wide band of frequencies is to be amplified, the load resistance must be low also. The tube must have a high transconductance.

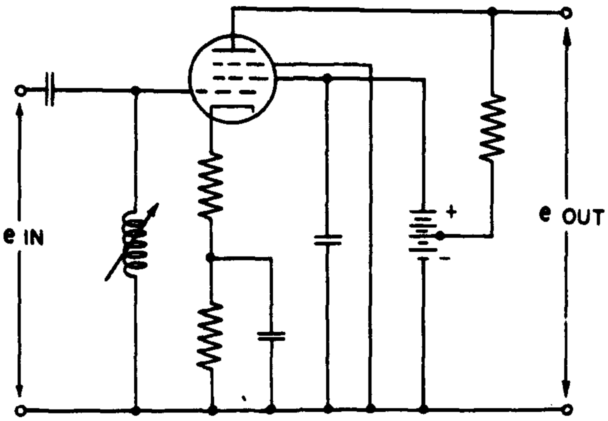


Figure 5-13.—Tuned amplifier for operation at radar frequencies.

The input capacitive reactance of a tube, such as the one in figure 5-13, varies inversely with the frequency of the a-c component of grid voltage, thus causing frequency distortion. Therefore, a part of the cathode resistance is unbypassed so that the correct amount of degeneration to overcome this distortion will be introduced.

Because of the low plate load, the plate supply voltage is also low, in this case lower even than the screen-grid voltage. The suppressor is returned to ground rather than to the cathode to increase the negative potential of the suppressor with respect to the plate.

Double tuning is seldom used because of the restrictions imposed by this type of amplifier on bandwidth.

VIDEO AMPLIFIERS

Video amplifiers such as those used in modern television sets are designed to give essentially uniform amplification of all frequencies from 30 cps to over 4 mc. Audio-frequency amplifiers, on the other hand, are considered good if they have a relatively flat response between 30 and 15,000 cps. Figure 5-14 compares the response of the two types of amplifiers.

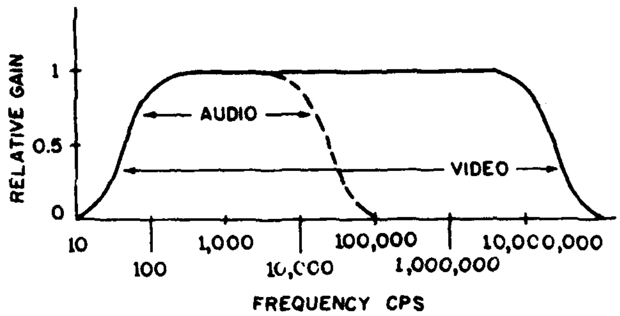


Figure 5-14.—Relative frequency response of audio and video amplifiers.

It has been shown that resistance coupling gives the best response over a wide range even at audio frequencies. This is especially significant in video amplifiers; however, to extend the range at both the low- and high-frequency ends, special compensation is necessary in video amplifiers.

Resistance-Capacitance Coupled Circuits

The *R-C* coupled amplifier circuit shown in figure 5-15

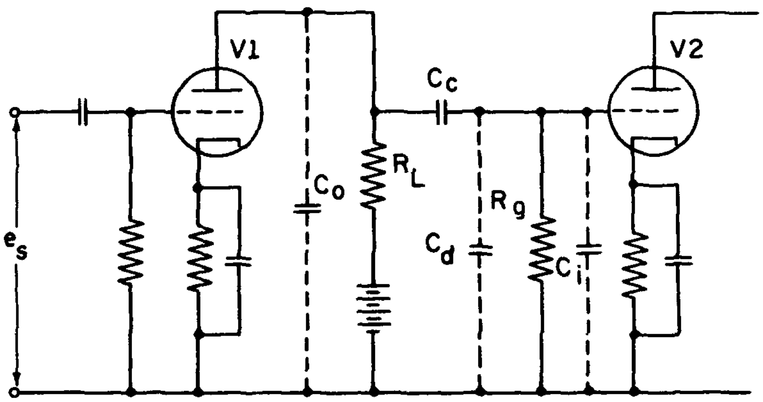


Figure 5-15.—*R-C* coupled amplifier.

indicates the effects that must be overcome if the range is to be extended on both ends of the frequency spectrum.

The high-frequency response is limited by the interelectrode output capacitance, C_o; the distributed wiring capacitance, C_d, and the input interelectrode capacitance, C_i. These three capacitances, acting in parallel, shunt the load, R_L, and reduce the output at the high frequencies.

The low-frequency response is limited by the reactance of the coupling capacitor, C_c. Thus at the lower frequencies the divider action of C_cR_g reduces the voltage available between the grid and cathode of $V2$.

It is obvious that the interelectrode capacitances of the tubes and the distributed capacitances of the wiring must be kept as low as practicable. Keeping the distributed capacitances low requires careful placement of the tubes in order in keep the leads as short as possible.

High-Frequency Compensation

There are various methods of extending the range of a video amplifier at the high-frequency end of the range, but perhaps the simplest and most effective is the shunt-peaked method, shown in figure 5–16, A. As mentioned previously, the gain at high frequencies is reduced because the load is shunted by C_o, C_d, and C_i. These same values can be made to extend the range if a small inductor, $L1$, is inserted in series with the load resistor, R_L, to form a parallel resonant circuit. If the value of $L1$ is properly chosen so that the circuit will be in resonance at the poitnt where the response curve begins to fall appreciably, the range can be extended. The value of $L1$ is critical. If the value is not correct, the amplification may be increased before the point at which the response curve begins to fall, with the result that frequency distortion ensues.

Series compensation may also be used to extend the range at the high-frequency end, as indicated in figure 5–16, B. In this instance an inductance, $L2$, of the proper value is added in series with the coupling capacitor, C_c, so that a

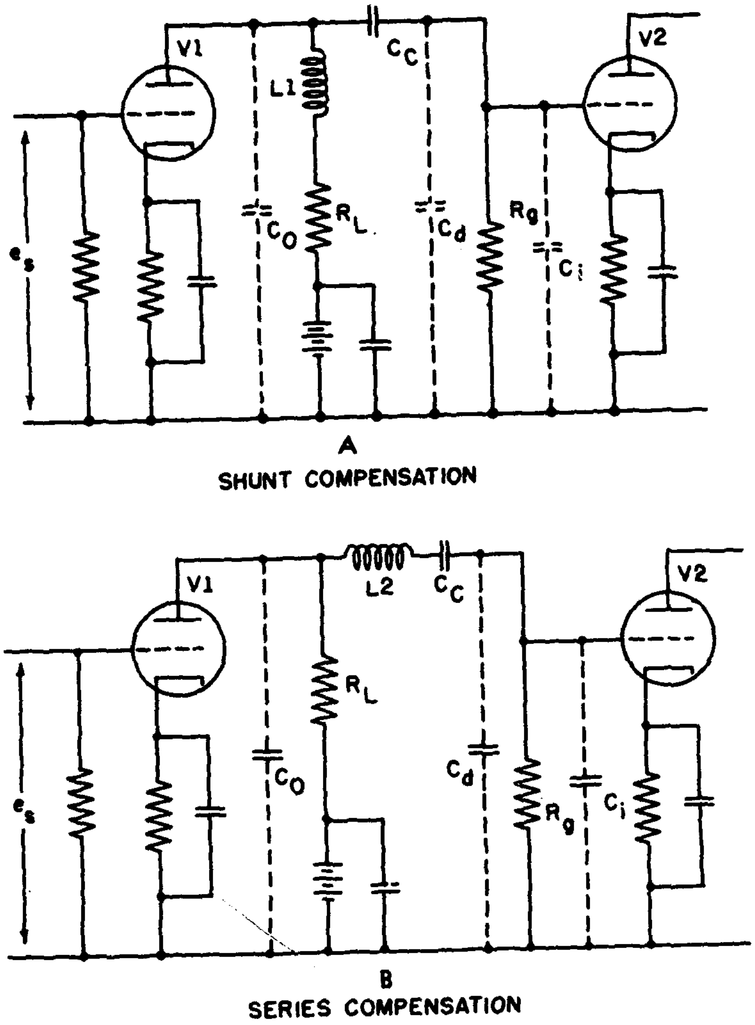


Figure 5-16.—High-frequency compensation.

series-resonant circuit is formed with the parallel combination of C_d and C_t. At resonance, increased current will flow through these capacitances and larger output voltage will be applied between the grid and cathode of $V2$.

The high-frequency peaking effect of shunt compensation in addition to the increased gain of series compensation may be obtained if both of these methods of compensation are used in the same coupling circuit. There are other factors, however, such as the transient response, which have to be considered in a network such as this.

Low-Frequency Compensation

At low frequencies, the distributed and interelectrode capacitances may be neglected but the reactance of the coupling capacitor becomes increasingly important. Since the reactance of this capacitor is

$$X_C=\frac{1}{2\pi fC_c},$$

it becomes appreciable at low frequencies. Consider a voltage divider made up of C_c and R_g (fig. 5–17) in which the

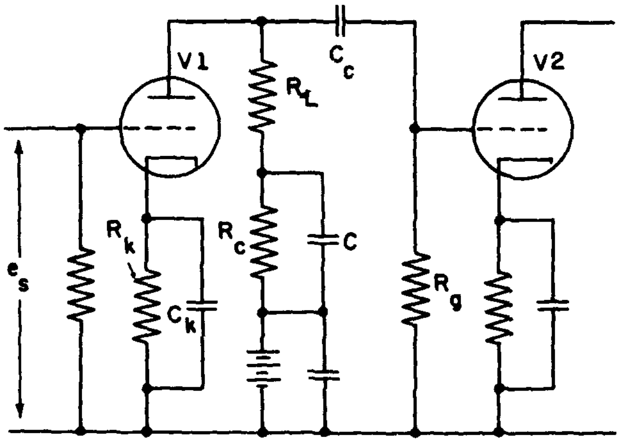


Figure 5–17.—Low-frequency compensation.

reactance of C_c is large, as it would be at low frequencies. More of the voltage would then appear across C_c and less would appear across R_g. Since in some practical applications the voltage gain must be maintained within 70 percent of the mid-frequency gain, this loss at low frequencies could not be tolerated.

The capacitance of C_c could be increased, but such a procedure would increase the stray capacitance and thus cut down the high-frequency gain.

Another factor that is perhaps more important than the loss of gain is the large shift in phase that occurs at these frequencies. If 10 stages of video amplification are used, a phase shift (lead) of about 2° is all that can be permitted for each stage.

The phase shift can be reduced by employing a large value of coupling capacitance and the largest permissible grid-leak resistance; but both of these expedients have their disadvantages also.

Amplifiers that do not have as critical requirements as video amplifiers may be made to operate satisfactorily by using large cathode and screen bypass capacitors and a coupling capacitor as large as practicable. Video amplifiers, however, require special compensation at the low frequencies.

Both the loss in gain and the increase in phase shift may be corrected by dividing the load resistance into two parts and bypassing one part with a capacitor. A circuit employing this method of low-frequency compensation is shown in figure 5–17.

The load resistance is made up of two parts, R_L and R_c, of which R_c is bypassed by C. At the higher frequencies the load is effectively R_L because R_c is bypassed by C which has a low reactance at these frequencies. At low frequencies, however, C offers a high impedance and the load is effectively R_L+R_c. This effective load increase causes a greater proportion of the plate signal to appear as output and thus counteracts the normal drop at the low frequencies. Care must be used in selecting the values of R_c and C so that uniform gain can be extended into the lower frequencies

beyond the point where the gain begins to fall off. Distortion will occur if the gain takes place before the curve begins to fall off.

Another component influencing low-frequency gain is the cathode-bypass capacitor. In order to prevent degeneration from occurring at the lower frequencies because of inadequate shunting, the capacitance of the cathode-bypass capacitor must be great enough to offer low impedance (with respect to the cathode resistor) at the lowest frequency to be amplified. Thus in video amplifiers the capacitances of cathode bypass capacitors are many times those of comparable audio amplifiers.

CATHODE FOLLOWERS

To achieve uniform response over a wide frequency range it has been shown that an amplifier should have a low effective input capacitance and a low effective load impedance. The over-all response may also be improved by the use of degenerative feedback. The cathode follower possesses these qualities, and in addition it may be used to match the impedance of one circuit to that of another.

The cathode follower is a single-stage class A degenerative amplifier the output of which appears across the unbypassed cathode resistor. The high input impedance (no grid current) and the low output impedance make it particularly useful for matching a high-impedance source to a low-impedance load. Thus, the cathode follower might be used between a pulse-generating stage and a transmission line whose effective shunt capacitance might be great enough to cause objectionable effects. More power, of course, can be delivered when the source is matched to the load. For example, a conventional amplifier having high output impedance would supply less power to a low-impedance coaxial line than would a cathode follower having an output impedance that corresponds to the load impedance.

The advantages obtained by the use of a cathode follower can be had only at the price of a voltage gain that is less

than unity; however, the circuit is capable of producing power gain.

As the name implies, the output voltage FOLLOWS the input voltage—that is, it has not only the same waveform but also the same instantaneous polarity (phase).

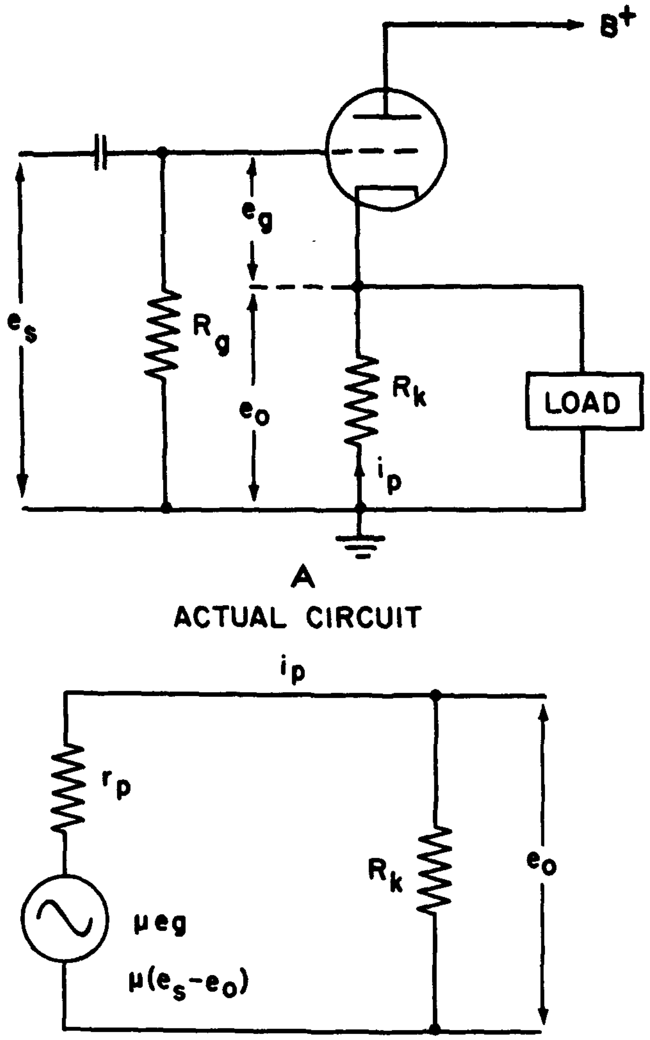
Circuit Operation

A conventional cathode follower is shown in figure 5–18. Under no-signal conditions a certain amount of plate current flows through R_k, and this flow establishes the normal bias. When a positive-going signal is applied to the grid, the plate current increases. This increase causes an increase in the voltage drop across R_k, giving the cathode a higher positive potential with respect to ground than it had under the no-signal condition. When a negative-going signal is applied to the grid, the opposite effect occurs. Thus, the output polarity FOLLOWS the polarity of the voltage applied between grid and ground.

Since R_k is not bypassed, degeneration occurs both on the positive half cycle when *plate current through* R_k increases the bias and on the negative half cycle when plate current through R_k decreases the bias. During the positive half cycle, the increase in bias subtracts from the input signal and reduces the amplitude of the grid-to-cathode voltage. Also during the negative half cycle, the bias adds to the input signal and the accompanying decrease in bias again reduces the amplitude of the grid-to-cathode voltage. Thus, in both half cycles the peak value of the a-c component of plate current is decreased and the output voltage is correspondingly reduced below the value it would have had if degeneration were not present.

Voltage Gain

To find an expression for the voltage gain (V. G.) of a cathode follower, use is made of the circuits shown in figure 5–18. The following may be established by inspection of part A of figure 5–18:



B
EQUIVALENT CIRCUIT– CONSTANT
VOLTAGE GENERATOR FORM

Figure 5-18.—Conventional cathode follower and equivalent circuit.

$$e_s=e_g+e_o,$$

$$e_o=e_s-e_g=i_pR_k,$$

and

$$e_g=e_s-e_o=e_s-i_pR_k;$$

also, from part B of figure 5-18,

$$i_p=\frac{\mu e_g}{r_p+R_k}=\frac{\mu(e_s-i_pR_k)}{r_p+R_k}.$$

It follows that

$$\mu(e_s-i_pR_k)=i_p(r_p+R_k),$$

and i_p may be solved as

$$i_p=\frac{\mu e_s}{r_p+(\mu+1)R_k}. \tag{5-23}$$

The output voltage is the product of i_p and R_k; and

$$e_o=\frac{\mu e_sR_k}{r_p+(\mu+1)R_k}.$$

Thus,

$$\text{V. G.}=\frac{e_o}{e_s}=\frac{\frac{\mu e_sR_k}{r_p+(\mu+1)R_k}}{e_s}=\frac{\mu R_k}{r_p+(\mu+1)R_k}.$$

For pentodes, μ is large and the term $\mu+1$ may be reduced to μ. The V. G. equation is then reduced to

$$\text{V. G.}=\frac{R_k}{\frac{r_p}{\mu}+R_k}=\frac{R_k}{\frac{1}{g_m}+R_k},$$

since

$$\frac{r_p}{\mu}=\frac{1}{g_m}.$$

It is seen that the denominator of the voltage gain fraction

is always greater than the numerator; hence, the gain is less than one.

Input Impedance

The input impedance of a cathode follower is high, and the effective input capacitance is low compared with that of a conventional amplifier. Both of these effects result from the degenerative action that occurs across the unbypassed cathode resistance.

Under no-signal conditions the grid is negative with respect to the cathode. When a positive-going signal is applied to the grid the bias is increased, because of degenerative action, to such an extent that no grid current will flow. The result is the same as if the input impedance had been increased. On the other half cycle when a negative-going signal is applied to the grid, the bias is decreased but no grid current can flow and the input impedance remains high.

The reduced input capacitance results from the fact that degeneration reduces the amplitude of the a-c component of the grid-to-cathode voltage, or in effect increases the input impedance, and thus causes less current to flow through the tube capacitances.

Because of the constant high impedance presented by the input to the cathode follower, it presents negligible loading to the circuit that drives it.

Output Impedance

A mathematical expression for the output impedance of a cathode follower is derived as follows: If the numerator and denominator of the right-hand side of equation (5–23) are divided by $\mu+1$, the resulting equation is

$$i_p = \frac{\frac{\mu e_s}{\mu+1}}{\frac{r_p}{\mu+1}+R_k}.$$

The tube now has an amplification factor of $\frac{\mu}{\mu+1}$ and an a-c plate resistance of $\frac{r_p}{\mu+1}$.

From the equivalent circuit of figure 5–18, B, the output impedance (with μe_g shorted) is the parallel combination of the a-c plate resistance $\frac{r_p}{\mu+1}$ and R_k. Therefore,

$$Z_o=\frac{R_k\frac{r_p}{\mu+1}}{R_k+\frac{r_p}{\mu+1}}=\frac{R_kr_p}{(\mu+1)\left[R_k+\frac{r_p}{\mu+1}\right]}=\frac{R_kr_p}{(\mu+1)\,R_k+r_p}.$$

The impedance is generally resistive; and if μ is very much greater than 1 the term $(\mu+1)\,R_k$ may be reduced to μR_k and the resulting equation reduced by dividing the numerator and denominator by r_p. Thus

$$Z_o=\frac{R_kr_p}{\mu R_k+r_p}=\frac{R_k}{\frac{\mu R_k}{r_p}+1}=\frac{R_k}{g_mR_k+1}.$$

The output impedance is low, and accordingly there is a minimum of amplitude distortion of the output signal even though current is drawn from the output terminals.

Distortion Caused by Limiting

Under normal operating conditions the output of a cathode-follower amplifier is practically free of amplitude distortion. However, if the input signal swings the grid voltage too far negative or positive, the output waveform will be limited or distorted in amplitude with respect to the input waveform. Beyond a certain negative value of grid voltage the plate current will be cut off and any further increase in negative grid potential will cause no corresponding change in plate current.

If the signal swings the grid voltage in a positive direction far enough for the grid to draw current, the loss in voltage

in the driving source limits the output signal and distortion again occurs.

The cathode-follower amplifier may be modified, as in figure 5-19, to adjust the grid bias to the correct value if

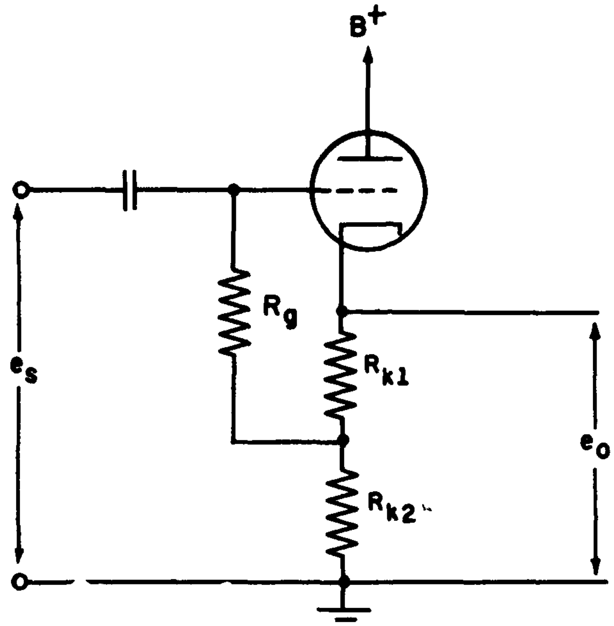


Figure 5-19.—Cathode follower modified to prevent limiting.

the cathode resistance is greater than the value required to give the correct grid bias and if limiting occurs only on the negative peaks of the input signal. In this modified circuit, the grid resistor, R_g, is connected to a point above ground on the cathode resistors R_{k1} and R_{k2}. This point is determined by the input voltage level. Thus the grid bias is reduced by an amount equal to the drop across R_{k2}.

Advantages of Cathode Followers

As previously stated one of the principal advantages of a cathode follower is that it can be used to match a high impedance to a low impedance. Thus it can take the volt-

age developed across a high impedance and supply a low impedance load with only a slightly less voltage but with a correspondingly large increase in current. One or more of the circuit elements of a cathode follower may be varied to achieve a more precise impedance match if the match is critical.

When tubes having a high mutual conductance are used, the low value of output impedance extends the amplification into the upper range of frequencies because the shunting effects of interelectrode and distributed capacitances are proportionately smaller. The low-frequency response is improved by allowing the d-c component of cathode current to flow in the load (fig. 5–18, A), thus avoiding the use of the series blocking capacitor.

The degenerative effect caused by the unbypassed cathode resistor increases the input impedance. Thus less shunting effect is offered to the previous stage, and a better over-all frequency response is produced.

As stated before, the input and output voltages have the same instantaneous polarity. When pulses are used it may be necessary to feed a positive- or a negative-going pulse to a load without polarity inversion. The cathode follower could thus serve two purposes—to prevent polarity inversion and to afford an impedance match.

Circuit stability is also improved, as in regular amplifiers, by degenerative feedback. Specifically, amplitude distortion occurring within the tube, the effect of plate-supply voltage variations, aging of tubes, production of harmonics, and other undesirable effects that occur within the stage are counteracted by this type of circuit.

However, these advantages are achieved at the expense of an over-all reduction in voltage gain. Normally, the voltage gain is slightly less than unity, but the circuit is capable of producing a gain in power.

PHASE INVERTERS

Since phase is generally associated with time, it is somewhat of a misnomer to apply this term to a device that

simply changes a positive-going signal to a negative-going signal or vice versa. In the case of a sine-wave signal, however, the effect is the same as if a 180° phase shift had occurred.

Paraphase amplifiers (phase splitters) produce, from a single input waveform, two output waveforms that have exactly opposite instantaneous polarities. If these two waveforms were produced as the result of a single sine-wave input they might be considered 180° out of phase, one waveform having been displaced 180° along the time axis.

One type of phase inverter is the transformer, with which the instantaneous polarity of the load may be reversed with respect to the source by reversing either the connections of the secondary leads to the load or the primary leads to the source. A conventional electron-tube amplifier (untuned and *R-C* coupled) also produces an output of opposite polarity to the input; and if no gain is desired, various methods may be employed to produce unity gain. Either single- or two-tube amplifiers may be used to convert one input waveform into two output waveforms of opposite polarity. Such amplifiers are called PHASE SPLITTERS or PARAPHASE amplifiers.

Transformer Phase Inverter

In operation, all transformers produce across the secondary an induced emf that is opposed to the change in flux producing it. The instantaneous polarity of the actual output voltage across a load depends on how the leads from the secondary are connected.

Figure 5–20 indicates phase inversion of square waves and sine waves. With square waves the polarity has simply been inverted. This is also true for sine waves but in this case it may be more convenient to refer to the inversion as a 180° phase shift—in effect, the same result as if the waveform had been moved along the time axis 180°. If no change in voltage is desired, a 1-to-1 turns ratio is employed.

A transformer with a center-tapped secondary or with a

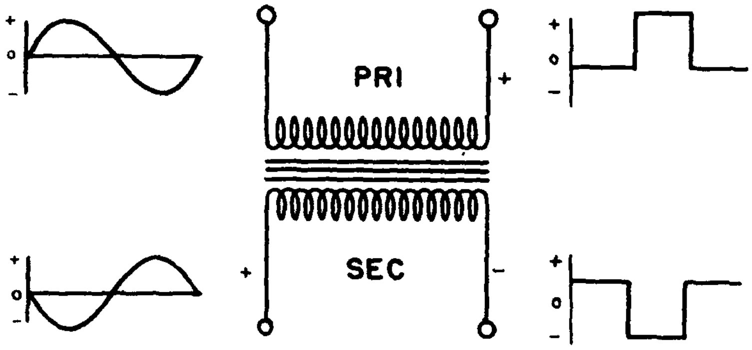


Figure 5-20.—Transformer phase inversion.

center-tapped resistor shunting the secondary, is used in class-B push-pull circuits to supply voltages of opposite instantaneous polarity to the grids of the tubes, as shown in figure 5-21.

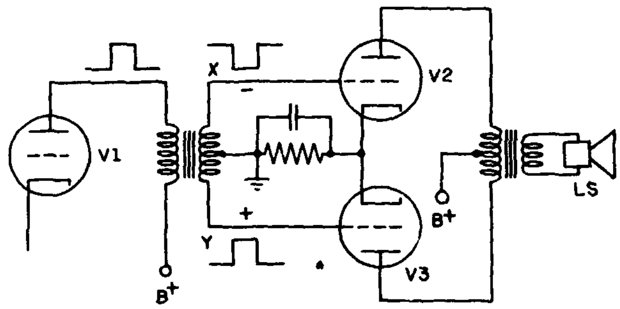


Figure 5-21.—Center-tapped transformer driving a push-pull amplifier.

If at a given instant the polarity of point X goes negative with respect to the grounded center tap, the polarity of point Y will go positive with respect to the center tap. Thus a negative potential is applied between the grid and ground of $V2$ and at the same time a positive potential is applied between the grid and ground of $V3$. This condition is necessary for the proper operation of a push-pull amplifier. Of course, the transformer must be tapped at the electrical center; otherwise the combined signal present in the output transformer will not be symmetrical with respect to the tap.

This type of transformer phase inverter has limited application because of distortions and losses inherent in transformers. For example, the loss in voltage through leakage reactance is greater for higher frequencies than it is for lower frequencies. The shunting capacitance effect and hysteresis losses also increase with frequency. Since in many circuits harmonics must be transmitted unattenuated and undistorted, the transformer phase inverter is generally replaced with a circuit that performs phase inversion without the use of transformers. The paraphase amplifier is such a circuit.

Electron-Tube Phase Inverter

As mentioned before, every electron tube used as a conventional amplifier introduces polarity inversion—that is, a negative-going signal between grid and ground causes a positive-going signal to be produced across the plate load. If there is to be polarity inversion with no gain in amplitude, some method must be employed to reduce the normal gain to unity. One method of reducing the normal gain is through the use of degenerative feedback. Degenerative feedback is readily obtained by omitting the cathode bypass capacitor.

Another method of reducing the gain is to employ a voltage divider in the input circuit. For example, if the normal gain of the tube is 100 the grid is tapped down on the divider so that one-hundredth of the available voltage is applied between grid and ground. If harmonics are to be included some method must be employed to reduce the input shunting effects of capacitance.

Single-Tube Paraphase Amplifier

One of the simplest forms of single-tube paraphase amplifiers is shown in figure 5–22. The values of resistors $R2$ and $R3$ are the same. Therefore the voltage drop across both of them is the same, since the same plate current flows through both. The instantaneous polarities, however, are

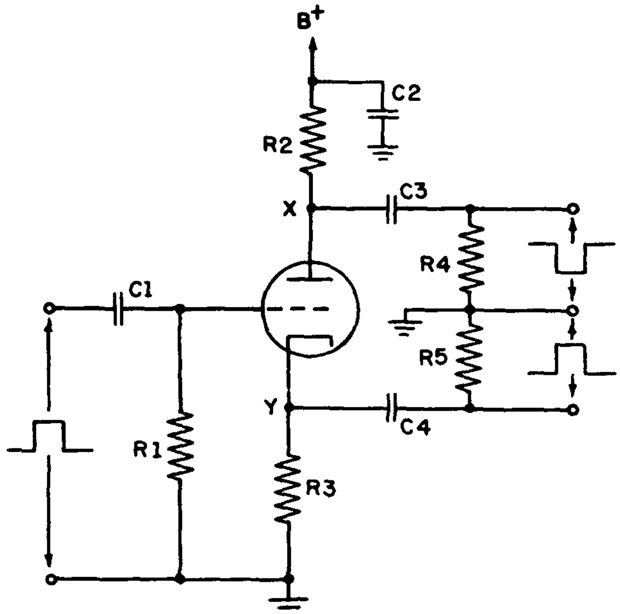


Figure 5-22.—Single-tube paraphase amplifier.

exactly opposite because at the instant a positive-going signal is applied to the grid, point *X* becomes less positive with respect to ground and point *Y* becomes more positive. These signals, with the polarities indicated in the figure, are impressed across load resistors *R*4 and *R*5 through blocking capacitors *C*3 and *C*4. *C*2 is the plate supply bypass capacitor.

In actual practice this basic type of single-tube paraphase amplifier may be modified to avoid some of the degenerative action due to the unbypassed cathode resistor or it may be compensated to permit a better frequency response.

Two-Tube Paraphase Amplifier

A 2-tube paraphase amplifier utilizes 1 tube as a regular amplifier and a second tube as a phase inverter, or these

functions may be performed by 2 sections of the same tube. The combination is frequently referred to as a PHASE INVERTER.

One of the simpler forms of two-tube paraphase amplifiers is shown in figure 5-23. *V*1 operates as a conventional

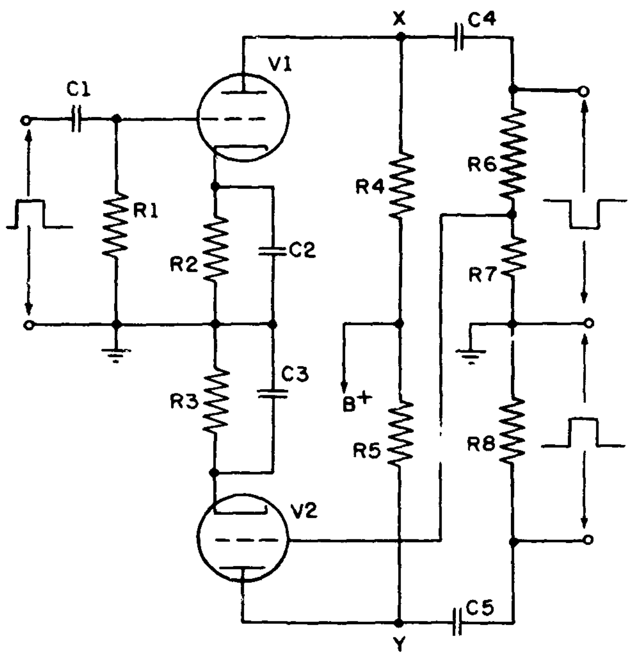


Figure 5-23.—Two-tube paraphrase amplifier.

amplifier having normal gain, and *V*2 operates as a phase inverter, the input of which is reduced to the same value as the input of *V*1. Thus *V*2 amplifies the signal as much as *V*1 and the output is essentially symmetrical about the zero-voltage reference line.

A positive-going signal on the grid of *V*1 causes an increase in plate current and a reduction in positive plate potential at point *X*. This reduction in positive potential is transmitted as a negative-going signal through coupling capacitor

$C4$ to resistors $R6$ and $R7$. The grid input to $V2$ is tapped down on resistors $R6$ and $R7$ to feed the proper magnitude of negative-going signal to $V2$. For example, if $V1$ and its associated circuit has a voltage gain of 50, the resistance of $R7$ should be one-fiftieth of the total value of $R6$ plus $R7$. At the instant a positive-going signal is applied to the grid of $V1$ a negative-going signal is thus applied to the grid of $V2$. The positive potential at point Y is increased, and a positive-going signal is applied to resistor $R8$, through coupling capacitor $C5$. At the same time the negative-going signal appears across resistors $R6$ and $R7$.

If the operating conditions of the two tubes are carefully chosen and the circuits are properly adjusted, the two amplified output signals should be essentially undistorted and of opposite instantaneous polarity. In actual practice this method presents some difficulty because the adjustments are critical. However, it is widely used as a means of driving class-A push-pull audio power amplifiers.

QUIZ

1. Name (a) three types of electromechanical loads and (b) two control functions that are applicable to d-c amplifier outputs.
2. What is the effect of positive feedback on amplifier gain and selectivity?
3. What is the effect of positive feedback on undesirable distortion introduced within the amplifier itself?
4. What is the effect of negative feedback on nonlinear distortion in an amplifier stage?
5. Why cannot distortion that is introduced in the first stage of an amplifier be eliminated by negative feedback applied across the last stage?
6. In which amplifier stages (high- or low-level) is feedback more effective?
7. If the high frequencies are to be amplified more than the low frequencies, what frequencies must be attenuated in the feedback network?
8. How does negative feedback affect the gain of an amplifier?
9. Negative feedback employing current feedback may be simply accomplished by leaving out which of the circuit elements?

10. What is the phase relation between the voltage feedback and the signal voltage in the V1 grid circuit of figure 5–7?
11. What are the relative values of circuit Q and g_m in a high-gain single-tuned transformer-coupled amplifier?
12. In a double-tuned transformer-coupled amplifier, how does the coupled impedance act with respect to the primary (in series or in parallel)?
13. How does the voltage gain vary with g_m, and ω in a double-tuned transformer-coupled amplifier?
14. How does an increase in the $\frac{L}{C}$ ratio of a double-tuned transformer-coupled amplifier affect the circuit Q?
15. What is the effect on a response curve (fig. 5–11, A) of a low coefficient of coupling?
16. What is the effect on a response curve (fig. 5–11, A) of increasing the coupling beyond optimum coupling?
17. Shunting a coil in a tuned circuit has what effect on the circuit Q?
18. In an amplifier modified for high-frequency operation, why is a part of the cathode resistance unbypassed?
19. What is the purpose of the coil when the shunt-peaked method of high-frequency compensation is used?
20. What is the purpose of dividing the load resistance into two parts and bypassing one part with a capacitor, as in figure 5–17?
21. How does the capacitance of the cathode bypass capacitor in figure 5–17 affect the low-frequency gain?
22. Is the relative voltage gain of a cathode follower greater or less than unity?
23. Why is a cathode follower so named?
24. What is the effect of degenerative action in the cathode follower on the peak positive grid-to-cathode voltage and on the peak negative grid-to-cathode voltage?
25. What is the relative magnitude of the output impedance of a cathode follower compared with the input impedance?
26. Omitting the series blocking capacitor in the cathode-follower output circuit affects the frequency response in what way?
27. What is the function of a paraphase amplifier?
28. What is the objection to using transformers as phase inverters in audio amplifiers?
29. Name two methods of reducing the gain in an electron-tube phase inverter.
30. In figure 5–23, why is the resistance of $R7$ less than that of $R6$?

CHAPTER 6

AUDIO POWER AMPLIFIERS

GENERAL

The primary function of the usual VOLTAGE amplifier is to increase the voltage of a signal to a higher value without distorting the waveform. Under ideal conditions no appreciable power is consumed from the preceding stage, and no appreciable power is supplied to the succeeding stage. In general, the output voltage is proportional to the product of the input voltage and the mu of the tube.

The primary function of a POWER amplifier is to deliver power to a load, and any increase in voltage is of secondary importance. Since the output power is proportional to the square of the grid voltage, the power amplifier must usually be preceded by one or more voltage amplifiers to raise the voltage to the proper level to operate the power stage.

Various types of tubes may be used as audio power amplifiers, including triodes (or tubes operated as triodes) and pentodes. The tubes may be operated singly as class-A amplifiers; or in pairs as in push-pull stages in which the tubes are operated as class-A, class-AB, or class-B amplifiers.

. general, audio power amplifiers have low amplification factors, low plate resistance, and high plate current. In order to obtain low plate resistance, the space between the plate and cathode is made smaller in a power tube than it is in a voltage, amplifier tube. Also, in a power tube the area of the plate is made larger and the cathode is designed to

supply a larger number of electrons. The grid must not block too many of the electrons flowing to the plate; accordingly the grid meshes are widely separated, and the amplification factor is thereby reduced.

Pentodes used as power amplifiers have higher amplification factors than triodes, but the plate resistance is proportionately higher.

Power amplifiers have numerous applications; the most familiar perhaps is the output stage of a radio receiver. Power is needed to operate the loudspeaker; therefore, the last audio stage is operated as a power amplifier.

CLASS-A TRIODE AMPLIFIERS

Class-A amplifiers are operated so that plate current flows during the entire input-voltage cycle. If the correct operating point, load impedance, and input voltage are chosen, the output waveform will be essentially the same as the input in all respects except for the amplitude.

Many limitations are imposed on class-A amplifiers if they are to be operated with minimum allowable distortion. The curvature of the lower portion of the i_p-e_g characteristic curve places a practical limit on the minimum current that may flow in the plate circuit. This limitation, in turn, places a limit on the negative swing of grid-signal voltage. If class-A_1 operation is assumed, the positive swing of the grid-signal voltage is limited by the magnitude of the bias in order that grid current may not flow.

Because of the need for minimum allowable distortion, maximum power output is limited. Maximum power output with minimum allowable distortion occurs for the triode when the load impedance is twice the plate resistance of the tube. This limitation, however, is not serious.

The efficiency of class-A amplifiers is limited to a low value (15 to 25 percent for class-A_1 operation) because appreciable d-c plate current flows during the entire grid-voltage cycle. Although its efficiency is low, the class-A amplifier has remarkably high fidelity if the proper operating conditions are chosen.

Load Line

One of the simplest methods of determining the output voltage and current components under a variety of operating conditions is by the use of a load line, as shown in figure 6–1, A. This line is a graph of the equation

$$e_p = E_B - i_p R_L,$$

where e_p is the instantaneous plate-to-cathode potential, E_B the plate supply voltage, and $i_p R_L$ the voltage drop across load resistor R_L.

The I_p-E_p curves across which the load line is plotted are known as STATIC CHARACTERISTIC CURVES because grid voltage changes and the accompanying plate current changes occur at constant plate potential. For example, if the grid bias is decreased from -35 to -30 volts at a constant plate potential of 250 volts the plate current will increase from 30 ma to 40 ma, or an increase of 10 ma.

In contrast to this action, DYNAMIC CHARACTERISTIC I_p-e_p CURVES take into account the change in plate voltage that always occurs with a change in plate current when a load resistor is connected in series with the plate. The load line makes possible the calculation of the dynamic characteristic. Thus if the grid bias is decreased from -35 volts (point B on the load line) to -30 volts, the plate current will increase from 30 ma to 33 ma, or an increase of 3 ma.

Hence, in the examples given the plate current increases 10 ma using the static curves, and only 3 ma using the dynamic characteristic because in the first case the plate voltage remains constant at 250 volts whereas in the second case the plate voltage decreases from 250 volts to 235 volts because of the increased voltage drop in load resistor R_L.

Load resistor R_L seldom exists as a real resistor but is used to represent the equivalent plate-load resistance in order to simplify the actual load circuit. For example, the actual load circuit might consist of a step-down transformer and its associated loudspeaker, in which case the actual B-supply voltage would be only large enough to supply the plate

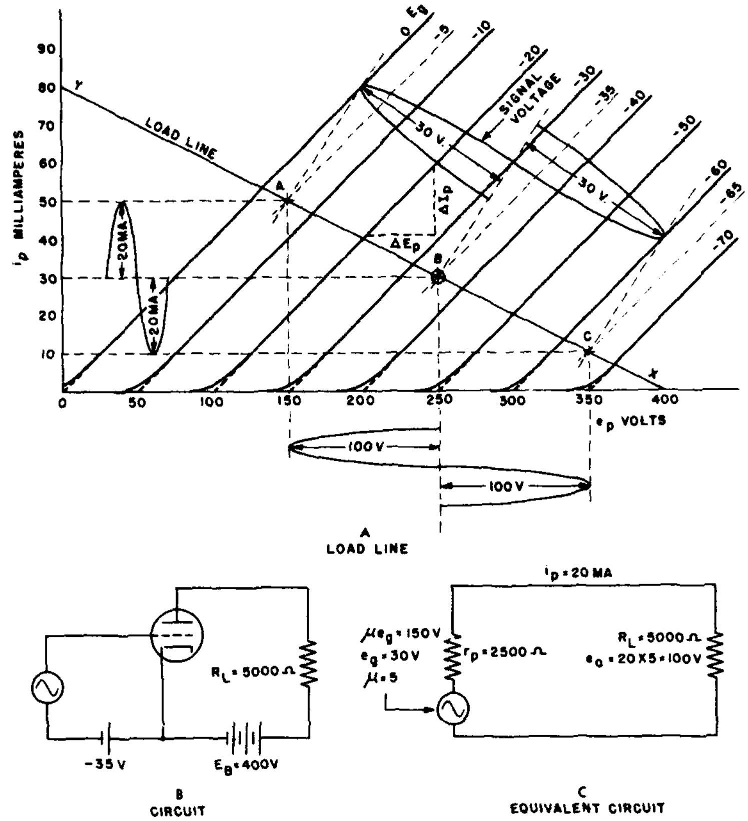


Figure 6–1.—Load line and circuits for a triode amplifier.

voltage plus the d-c voltage drop in the transformer primary. However, the load line is based upon the theoretical B-supply voltage that would be required if R_L were actually a load resistor.

The terminal points Y and X on the Y axis and the X axis, respectively, are easily established. Thus if the electron tube resistance could be reduced to zero, the plate current would become a maximum value and the B-supply voltage

would appear across the load resistance. The plate current would be

$$i_p = \frac{E_B}{R_L}.$$

This value of i_p determines the Y-axis terminal of the load line, indicated in figure 6–1, A, as point Y.

When no plate current flows, the full value of the plate supply voltage is applied between the plate and cathode because there is no voltage drop across R_L. Thus,

$$e_p = E_B.$$

This value of e_p determines the X-axis terminal of the load line and is indicated as point X.

When a number of i_p-e_p curves for various values of grid bias are included on the graph, the plate current and voltage corresponding to a given bias can be determined at a glance. Thus in figure 6–1, A, point A represents a grid bias of -5 volts. A line drawn horizontally from this point to the Y axis indicates a plate current of approximately 50 milliamperes. Likewise a line drawn vertically downward to the X axis from point A indicates a corresponding plate voltage of approximately 150 volts.

A line drawn horizontally from point B (the operating point where the i_p-e_p curve for a grid bias of -35 volts intersects the load line) to the Y axis indicates a plate current of 30 milliamperes. The corresponding plate voltage is approximately 250 volts. The same procedure is followed in determining the plate current and voltage for point C or any other given value of grid bias.

Point B is called the OPERATING POINT because it represents the grid bias when no signal is applied. It is therefore the point about which the grid voltage will vary when a signal is applied. The instantaneous values of plate current and voltage may be considered as being determined by the position of an imaginary point oscillating about B along the load line. The limits of the oscillations are determined by

the point at which the grid begins to draw current on the positive swing and by the point at which the load line extends into the nonlinear portion of the plate-current curves on the negative swing. In establishing these limits it is of course assumed that the source is capable of supplying the input voltage swing between these limits.

If, as in figure 6–1, A, the grid voltage is assumed to have a maximum positive swing of 30 volts, then at the positive peak of the swing the net voltage impressed on the grid is $30-35=-5$ volts. The plate current at this instant reaches its maximum value of 50 milliamperes, and the plate voltage reaches its minimum value of 150 volts. At the negative peak of the input signal the total voltage between grid and cathode is $-30-35=-65$ volts. The plate current at this instant reaches its minimum value of 10 milliamperes and the plate voltage reaches its maximum value of 350 volts.

From the maximum to minimum values the peak a-c plate voltage is

$$e_p=\frac{E_{max}-E_{min}}{2}=\frac{350-150}{2}=100 \text{ v.}$$

The peak a-c plate current is

$$i_p=\frac{I_{max}-I_{min}}{2}=\frac{50-10}{2}=20 \text{ ma.}$$

By the use of the I_p-E_p curves of figure 6–1, A, the static characteristics of the triode may be determined. The plate resistance, for example, is the ratio of the change in plate voltage to the corresponding change in plate current for a constant grid bias—that is

$$r_p=\frac{\Delta E_p}{\Delta I_p}=\frac{250-200}{60-40}=\frac{50}{20}=2.5 \text{ k-ohms.}$$

The mu of the triode is the ratio of the change in plate voltage necessary to produce a certain change in plate current, to the change in grid voltage necessary to produce the same change in plate current. Thus,

$$\mu=\frac{\Delta E_p}{\Delta E_g}=\frac{250-200}{30-20}=\frac{50}{10}=5.$$

The mutual conductance, g_m, of the triode is the ratio of the change in plate current to the change in grid voltage producing it. If the change in plate current is expressed in microamperes and the change in grid signal is expressed in volts, the transconductance will be expressed in micromhos. Thus,

$$g_m=\frac{\Delta I_p}{\Delta E_g}=\frac{(60-40)10^3}{30-20}=\frac{20\times10^3}{10}=2{,}000 \text{ micromhos.}$$

In terms of μ and r_p the mutual conductance may be expressed as

$$g_m=\frac{\mu}{r_p}=\frac{5.0}{2{,}500}=0.002 \text{ mhos, or } 2{,}000 \text{ micromhos.}$$

The voltage gain of an amplifier stage is the ratio of the signal-voltage output to the signal-voltage input—

$$\text{voltage gain}=\frac{e_o}{e_g}.$$

Under the conditions established in figure 6–1, A, the voltage gain is

$$\text{V. G.}=\frac{e_o}{e_g}=\frac{100}{30}=3.33.$$

From the equivalent circuit of figure 6–1, C, μ is 5, r_p is 2.5 k-ohms, and R_L is 5 k-ohms. Expressed in terms of these values, the voltage gain is

$$\text{V. G.}=\frac{\mu R_L}{r_p+R_L}=\frac{5\times5}{2.5+5}=3.33.$$

Power Output

The power output of the amplifier of figure 6–1, B, may be calculated from the I_p-E_p curves of figure 6–1, A. The

plate voltage swings between 150 v and 350 v as the plate current swings from 50 ma to 10 ma. The peak a-c plate current is $\frac{50-10}{2}=20$ ma. The maximum a-c signal voltage is $\frac{350-150}{2}=100$ volts. The power output of the triode amplifier is

$$P_o=\frac{E_{max}I_{max}}{2}=\frac{100\times 0.02}{2}=1 \text{ watt.}$$

The power output may also be calculated from the equivalent circuit of figure 6-1, C, as the power delivered to the load resistor, R_L.

The voltage output of an amplifier stage is proportional to μe_g and the power output is proportional to $(\mu e_g)^2$.

Since μe_g acts in series with r_p and R_L in the equivalent circuit, the a-c component of plate current is

$$i_p=\frac{\mu e_g}{r_p+R_L}.$$

The output voltage appears across R_L as i_pR_L. Thus

$$e_o=i_pR_L=\frac{\mu e_g R_L}{r_p+R_L}.$$

The output power in watts, when I_p and E_o are effective (rms) values, becomes

$$P_o=I_pE_o=\frac{(\mu E_g)^2R_L}{(r_p+R_L)^2}. \tag{6-1}$$

In the example in figure 6-1, C, R_L is 5 k-ohms, r_p is 2.5 k-ohms, E_g is 21.2 volts (rms), and μ is 5. The power output is

$$P_o=\frac{(5\times 21.2)^2\times 5\times 10^3}{(2.5\times 10^3+5\times 10^3)^2}=1 \text{ watt.}$$

Since maximum transfer of energy occurs when $R_L=r_p$, the power output for this condition becomes

$$P_o=\frac{(\mu E_g)^2 r_p}{(2r_p)^2}=\frac{(\mu E_g)^2}{4r_p}. \qquad (6\text{-}2)$$

In the example in figure 6–1, C, if R_L is changed to 2.5 k-ohms, the power output becomes

$$P_o=\frac{(5\times 21.2)^2}{4\times 2.5\times 10^3}=1.12 \text{ watts.}$$

The problem of distortion is present in the power amplifier the same as it is in the voltage amplifier and here again a balance must be struck between maximum power output and minimum allowable distortion. Since the human ear is not particularly sensitive to distortion below about 5 percent, this much may be allowed in the output circuit. When the term "undistorted" is used in the following considerations, distortion up to 5 percent is allowed. Experiments show that when $R_L=2r_p$ the most noticeable distortion (that due to the second harmonic) is reduced to less than 5 percent. The reduction in power output, when $R_L=2r_p$ as compared with the power output when $R_L=r_p$, is only about 11 percent, as shown by the examples in figure 6–1, C.

Increasing the load resistance in the plate circuit of an electron-tube amplifier tends to reduce the slope of the i_p-e_g characteristic curve, as shown in figure 6–2. The curves are flattened because the higher the load resistance the lower is the voltage that is available at the plate, and consequently the lower becomes the plate current.

In general, the curve will have an appreciable bend if the load impedance is equal to the plate resistance of the power-amplifier tube. Unfortunately this value of load impedance would permit maximum transfer of power.

As mentioned previously, it has been found experimentally that MAXIMUM UNDISTORTED POWER OUTPUT may be achieved when the load impedance is approximately twice the plate resistance of the tube and the plate current variations are at the maximum permissible value for class-A operation.

The equation for maximum undistorted power output when

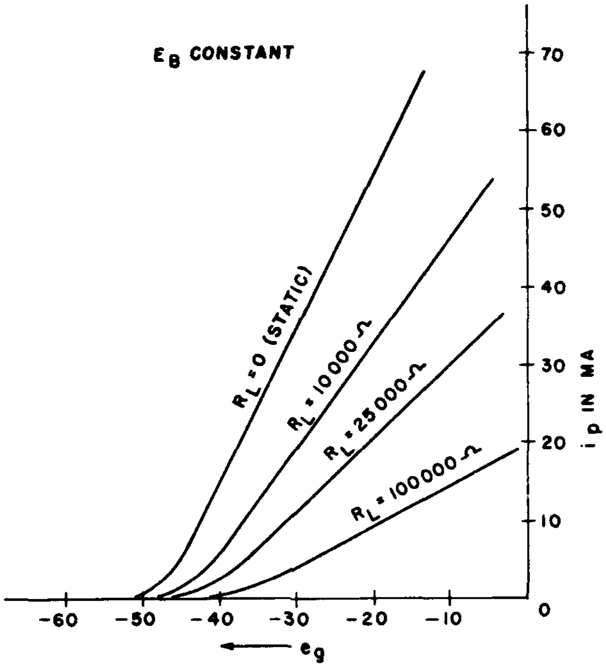


Figure 6–2.—i_p-e_g curves for a given plate voltage under various load conditions.

the load impedance, $R_L=2r_p$, is established as follows: In equation (6–1) if R_L is replaced by $2r_p$ and E_g is expressed in effective (rms) volts,

$$P_o=\frac{(\mu E_g)^2 2r_p}{(r_p+2r_p)^2}=\frac{(\mu E_g)^2}{4.5\ r_p}. \qquad (6\text{–}3)$$

A comparison of equations (6–2) and (6–3) verifies the reduction in power output as a result of making the load impedance twice the plate resistance to be only about 11 percent. This reduction is in agreement with the examples in figure 6–1, C.

Second-Harmonic Distortion

A careful examination of figures 6-2 and 6-3 will reveal some of the problems involved in designing triode power amplifiers and also how the concept of second-harmonic addition may be used to describe distortion.

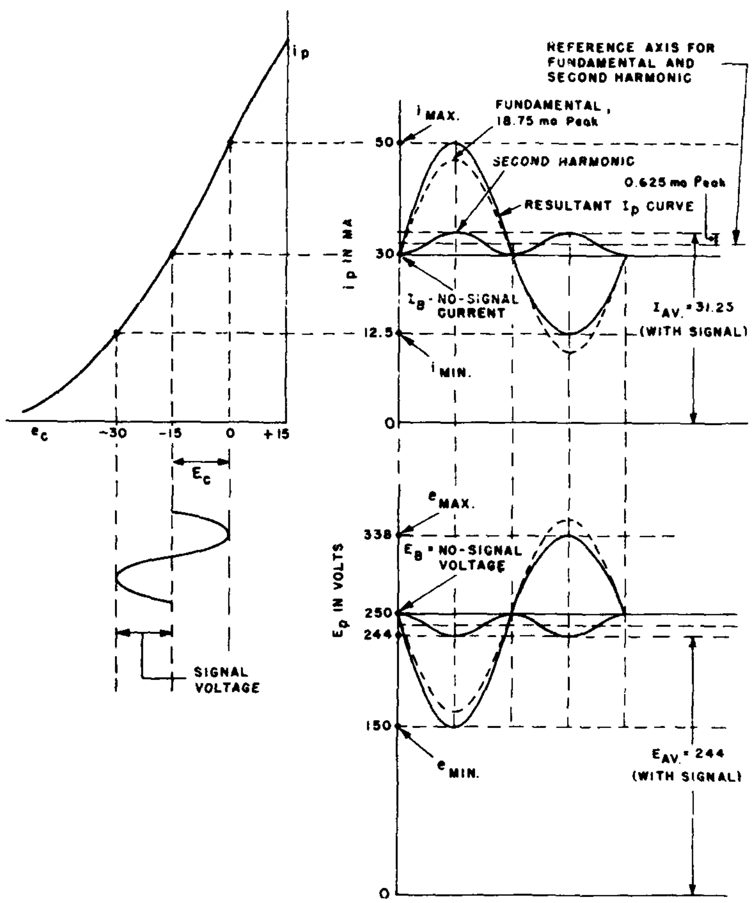


Figure 6-3.—Second-harmonic distortion in a class-A power amplifier.

If the bias is adjusted so that only the straight portion of the i_p-e_g characteristic curve is used, the signal-voltage swing will be greatly reduced. Since the power output varies as the square of the input-signal voltage, there is a practical limit to the amount of restriction that can be imposed. The load impedance may be further increased, but here again the current variation in the plate circuit will be reduced accordingly, and so will the power. If the bias is reduced, grid clipping may occur and second harmonics, as well as other harmonics, may be introduced. Harmonics are also introduced if the bias is so high that the negative half of the plate-current swing is reduced. If the signal voltage is too high, both grid clipping and operation beyond the bend in the lower portion of the i_p-e_g characteristic curve may occur.

When certain types of distortion are referred to as second- or third-harmonic distortion or perhaps as even- or odd-harmonic distortion, these harmonics are not necessarily actually produced in the tube. Rather the EFFECT on the output signal is the same as if these harmonics had been introduced. This concept is useful in explaining many types of distortion caused by electron tubes.

A consideration of the distortion of the plate-current curve in figure 6–3 reveals the presence of a second-harmonic component (deliberately exaggerated in the figure). In this instance the positive half of the plate-current curve is increased beyond the proportions of a sine curve and the negative half is decreased.

The lack of symmetry in the plate-current curve may be adequately explained if a second-harmonic signal is assumed to exist, as shown above the no-signal current line.

It is helpful to make a brief mathematical analysis of the contributions made by the second-harmonic component, and by the fundamental, to the resultant plate current.

The AVERAGE PLATE CURRENT, I_{av}, may be established by inspection—

$$I_{av}=\frac{I_{max}+I_{min}}{2}=\frac{50+12.5}{2}=31.25 \text{ ma.}$$

The MAXIMUM PLATE-CURRENT SWING of the second-harmonic component, I_2, is

$$I_{av}-I_B=31.25-30=1.25 \text{ ma},$$

where I_B is equal to the no-signal value of plate current. The maximum value of the second-harmonic current is

$$I_{2\,max}=\frac{1.25}{2}=0.625.$$

Also, the maximum plate-current swing of the fundamental is

$$I_1=I_{max}-I_{min}=50-12.5=37.5,$$

and the maximum value of the fundamental is therefore $\frac{37.5}{2}=18.75$ ma.

The equation for the AVERAGE VALUE of the second-harmonic component of plate current is

$$I_2=\frac{I_{max}+I_{min}-2I_B}{4}.$$

With the aid of the voltage curve, the contributio of the two components to the plate-voltage variations may be determined in a similar manner.

The average power output, P_o, is

$$P_o=\frac{(I_{max}-I_{min})(E_{max}-E_{min})}{8}$$

$$=\frac{(0.05-0.0125)(338-150)}{8}=0.881 \text{ watts}.$$

The percentage of distortion (P. D.) due to the second-harmonic current component is determined by dividing the second-harmonic component by the fundamental component and multiplying the result by 100. Thus,

$$P.D.=\frac{I_2}{I_1}\times 100=\frac{0.625}{18.75}\times 100=3.3 \text{ percent.}$$

The general equation for determining the percentage of distortion due to the second harmonic is

$$P.D.=\frac{I_{max}+I_{min}-2I_B}{2(I_{max}-I_{min})}\times 100.$$

Output Transformer

The output transformer used with audio power amplifiers serves as an IMPEDANCE-MATCHING device. Since the plate resistance of a power-amplifier tube may range from perhaps 1,000 ohms to more than 20,000 ohms, and since the impedance of the loudspeaker voice coil may range down to 4 ohms, the output transformer has a step-down turns ratio to provide the correct ratio of primary voltage and current to secondary voltage and current.

Impedance matching by means of a transformer is illustrated in figure 6–4. It is recalled that the output voltage of

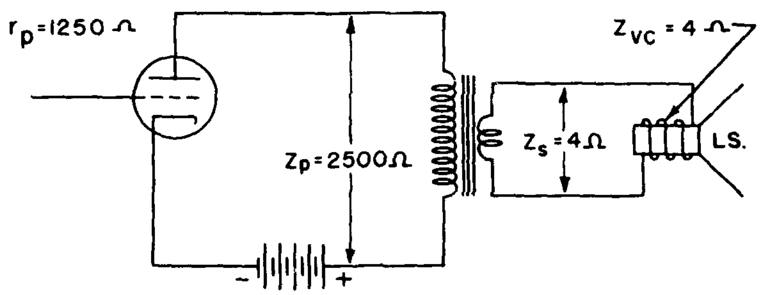


Figure 6–4.—Output transformer used as an impedance-matching device.

a transformer varies directly as the turns ratio, and that the output current varies inversely as the turns ratio—that is,

$$\frac{E_p}{E_s}=\frac{N_p}{N_s}$$

and

$$\frac{I_s}{I_p}=\frac{N_p}{N_s}.$$

If the 2 left sides and the 2 right sides of these equations are multiplied together, the impedance may be determined. Thus,

$$\frac{E_p}{I_p}\times\frac{I_s}{E_s}=\frac{N_p^2}{N_s^2}.$$

The primary impedance of a matching transformer is defined as the ratio of rated primary voltage to rated primary current. Similarly, the secondary impedance is the ratio of rated secondary volts to rated secondary amperes. If Z_p is substituted for $\frac{E_p}{I_p}$ and $\frac{1}{Z_s}$ is substituted for $\frac{I_s}{E_s}$, the impedance equation becomes

$$\frac{Z_p}{Z_s}=\left(\frac{N_p}{N_s}\right)^2.$$

Thus, the ratio of the two impedances that a transformer can match is equal to the turns ratio squared.

As a practical illustration, find the turns ratio needed for the transformer shown in figure 6–4. Since the plate resistance is 1,250 ohms, the primary impedance is given as twice this value, to permit maximum undistorted power output. The power fed to the 4-ohm voice coil, however, will be reduced unless the proper impedance is afforded by the transformer. The turns ratio that will satisfy this condition is readily determined by extracting the square root of both sides of the last equation. Thus,

$$\frac{N_p}{N_s}=\sqrt{\frac{2{,}500}{4}}=25.$$

The amount of power that can be handled by an output transformer is determined by the current and voltage RATINGS of the windings and the allowable losses. The primary frequently contains a d-c component that limits the

incremental inductance and frequency response. The equation for the induced voltage, E (rms), of a transformer winding is

$$E=4.44\, fNBA\, 10^{-8},$$

where f is the frequency in cycles per second, N the number of turns in the winding, B the flux density of the core, and A the cross-sectional area of the core. In a given transformer the induced voltage is proportional to the product of the frequency and the flux density. At low frequencies the flux density is high and more distortion is introduced because of the saturation of the iron. The maximum allowable flux density is determined by the maximum allowable distortion.

The output transformer causes a reduction in the output of a power amplifier at both the high and low frequencies. The reduced output at the low frequencies results from the shunting action of the transformer primary inductance on the load, as indicated in figure 6–5, A.

The following symbols are used in figure 6–5:

Symbol	*Definition*
μ	Amplification factor of tube
E_g	Input voltage (rms)
r_p	Plate resistance
R_1	Resistance of primary winding
L_1	Leakage inductance of primary
L_p	Incremental primary inductance
N	Primary-to-secondary turns ratio
L_2	Leakage inductance of secondary
R_2	Resistance of secondary winding
R_L	Load resistance
E_L	Voltage (rms) developed across the load
k	Coefficient of coupling

The middle-frequency gain is independent of frequency, as indicated by the absence of reactance in the equivalent circuit (fig. 6–5, B). The reactance of the primary inductance is large enough for its shunting effect to be disregarded and the leakage reactances are low enough to be neglected.

The reduced output at the high frequencies (fig. 6–5, C) results from the loss in voltage through the leakage reactances

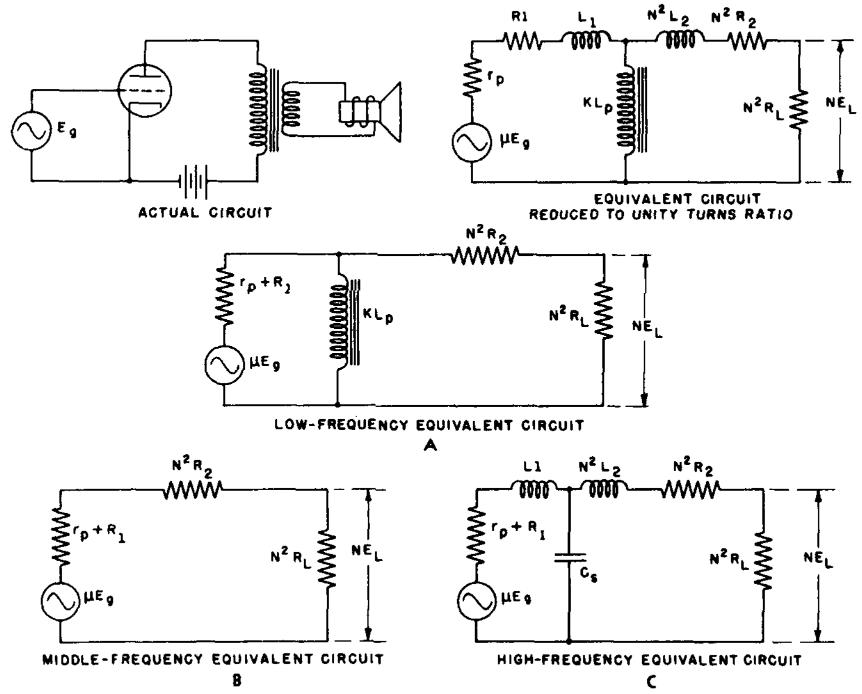


Figure 6-5.—Circuit analysis of an output transformer.

as a result of (1) load current and (2) capacitive current due to shunting capacitance. A frequency-response curve illustrating the reduction in gain at the low and high frequencies is shown in figure 6–6.

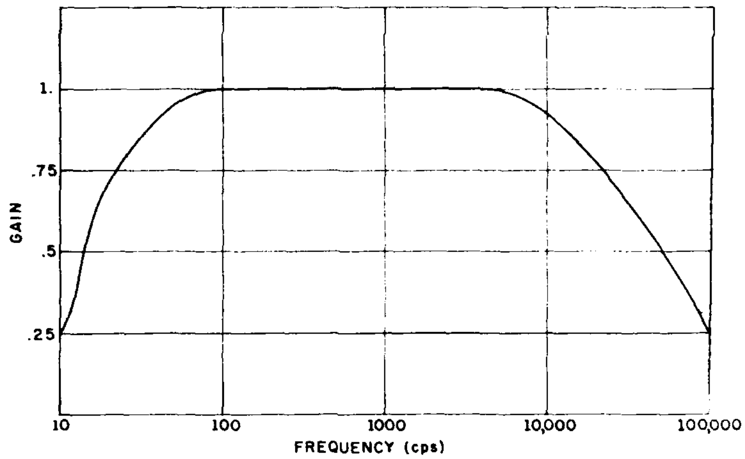


Figure 6–6.—Frequency response curve of an output transformer.

In order to extend the flat portion of the frequency-response curve into the low-frequency region, the transformer should have a high primary inductance. In order to extend the flat portion into the high-frequency region, the leakage inductance should be low. With a given triode and transformer, increasing the load resistance improves the high-frequency response without disturbing the low-frequency response very much.

The voltage across the load at the various frequency ranges may be determined from the equivalent circuit shown in figure 6–5. For example, at the middle range of frequencies,

$$NE_L=\frac{\mu E_g N^2 R_L}{(r_p+R_1)+N^2(R_2+R_L)},$$

and

$$E_L = \frac{\mu E_g N R_L}{(r_p + R_1) + N^2 (R_2 + R_L)}.$$

PUSH-PULL POWER AMPLIFIERS

A number of advantages are to be gained by the use of a push-pull amplifier as the output stage of an audio-frequency amplifier. Second harmonics and all even-numbered harmonics, as well as even-order combinations of frequencies, will be effectively eliminated if the tubes are properly balanced and if the frequencies are introduced within the output tubes themselves.

Hum from the plate power supply, which may be present in the single-tube amplifier, is substantially reduced in the push-pull amplifier because ripple components in the two halves of the primary transformer are in phase and tend to counteract each other in the output.

Plate-current flow through the two halves of the primary winding is equal and in opposite directions. Therefore there is no d-c core saturation and the low-frequency response is improved.

Regeneration is also eliminated because signal currents do not flow through the plate-voltage supply when the circuit is operated as a class-A amplifier.

The last voltage amplifier preceding the push-pull power amplifier stage may be either resistance- or transformer-coupled to the power stage. If the power amplifier is operated class A or class AB, the driver commonly employs resistance coupling because it affords a better frequency response. A phase-inverter tube, or section of a tube, must be used in connection with the resistance-coupled driver to provide the correct phase relation at the input of the push-pull stage.

When the power tubes are operated class B, an input transformer employing a step-down turns ratio is commonly used. The transformer not only supplies the grid current

necessary for class-B operation, but at the same time permits an instantaneous signal voltage of the correct polarity to be applied to the grids of the two power tubes.

Class-B power amplifiers draw practically no plate current when no signal is applied, and their plate efficiency is much higher than that of class-A amplifiers. They are subject, however, to third-harmonic distortion and the operating conditions are critical enough to require the supervision of trained personnel. These amplifiers are therefore used in transmitters but not generally in home installations.

Transformer-Coupled Push-Pull Amplifier

A transformer-coupled push-pull amplifier is shown in figure 6–7. It is assumed that the following operating con-

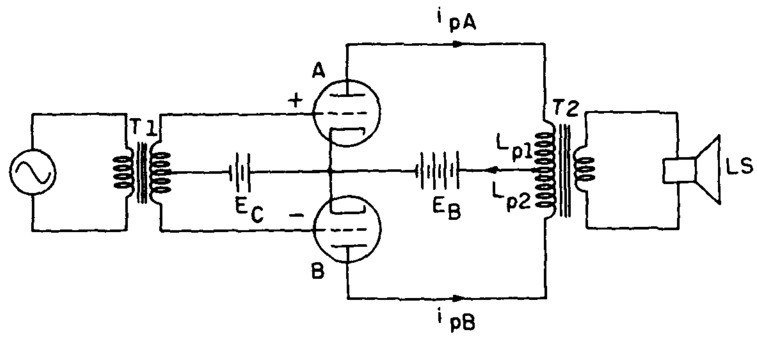


Figure 6–7.—Transformer-coupled push-pull amplifier.

ditions are in effect: The grid bias is such that the plate current in each tube flows during the entire input-voltage cycle (class A), and during the positive half of the cycle the grid-voltage excursion extends over a portion of the lower bend of the i_p-e_g characteristic curve. The tubes are also properly matched and operate into the correct loads.

When no signal voltage is applied, equal plate currents flow through each tube. Equal currents also flow through each half of the primary of the output transformer toward the center tap. The magnetomotive forces resulting from the

currents are equal and opposite and therefore cancel, leaving no magnetic field due to the d-c component of the plate current. This cancellation effect is a big advantage over the single-tube output in which direct current flows continuously through the primary winding and establishes a d-c field component.

A signal voltage across the secondary of the input transformer, $T1$, will at a given instant have polarities as indicated. This voltage will be divided equally between tubes A and B. The push-pull arrangement thus requires, and will handle, twice the input voltage of a single tube under similar operating conditions. The grid of tube A is positive with respect to the center tap at the instant the grid of tube B is negative. Plate current increases in tube A and decreases a proportionate amount in tube B.

The increase in current flowing down through L_{p1} and the decrease in current flowing up through L_{p2} constitute 2 magnetomotive forces that combine additively to produce an output voltage in the secondary that is proportional to the sum of these 2 components.

Second harmonics are eliminated in the push-pull output, as shown in figure 6–8. The dynamic i_p-e_g curve for tube B is inverted with respect to that of tube A in order to show the phase relation between the signal components of the two tubes. Thus, when the input signal swings the grid voltage of tube A in a positive direction it is swinging the grid voltage of tube B the same amount in a negative direction. Plate current in tube A increases as plate current in tube B decreases. The plate current swing about the X-X axis for tube A is not symmetrical because the tube is operating on a nonlinear portion of the i_p-e_g characteristic curve. The same condition is true of the plate current swing about the X'-X' axis for tube B.

The plate-current curves of each tube may be resolved into a fundamental and second harmonic. Thus, the axis of the fundamental and its second harmonic is displaced from the axis of the original plate current curve by an amount equal to the peak value of the second harmonic. Combining

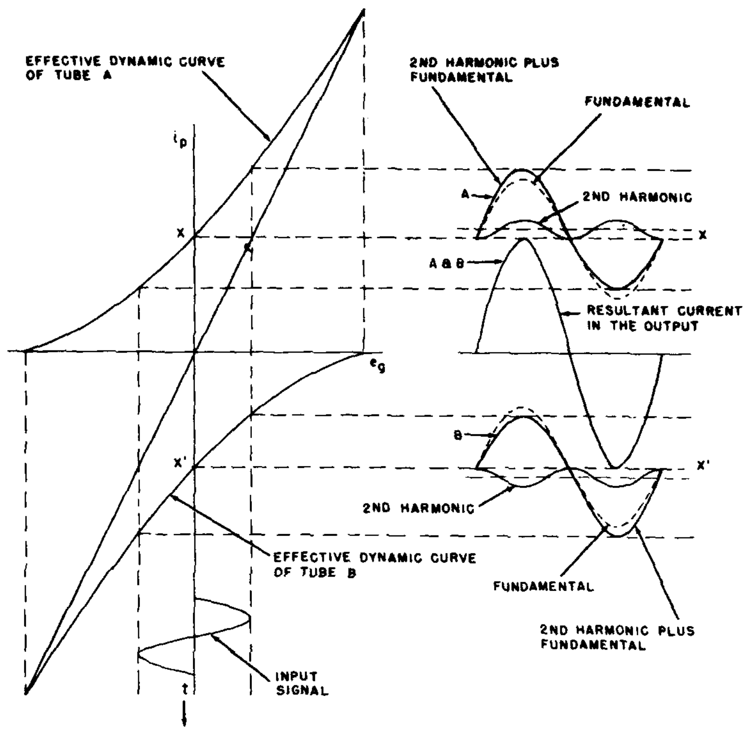


Figure 6-8.—Graph showing second-harmonic elimination.

the fundamental components of both tubes gives an output of twice the amplitude of one tube. However, when the second harmonics are combined, the resultant is zero because they are 180° out of phase. The fundamental output current has the same waveform as the input voltage—an effect that would have been produced had both tubes been free of second harmonics.

Power Output

The power output of a class-A push-pull amplifier is conveniently determined by the use of the equivalent circuit

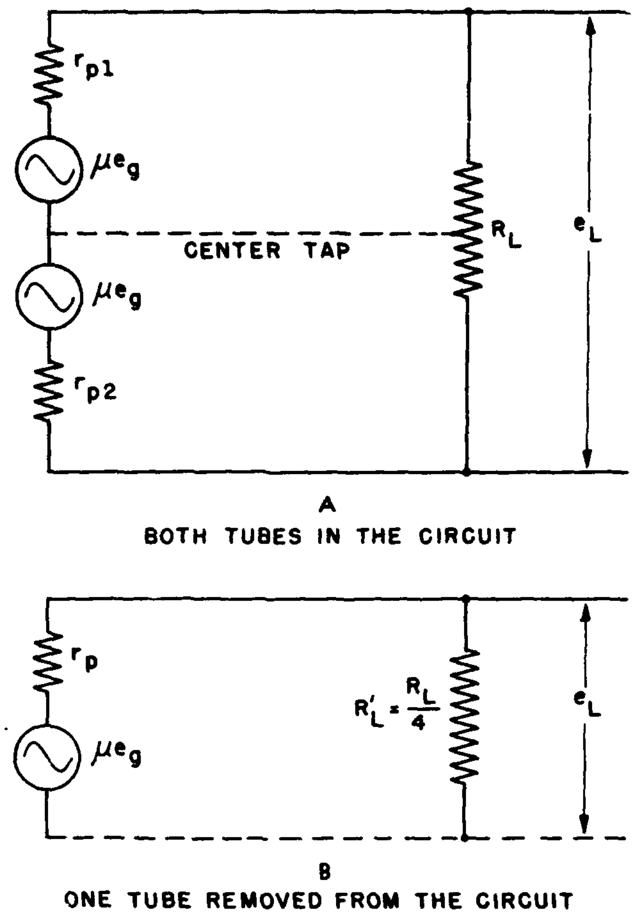


Figure 6-9.—Equivalent circuit of a push-pull amplifier.

shown in figure 6-9, A. If e_g is the maximum input-signal voltage then the maximum current through the load is

$$I_{max}=\frac{2\mu e_g}{r_{p1}+r_{p2}+R_L},$$

and the peak voltage across the load is

$$E_{max}=R_L I_{max}$$

$$=\frac{2R_L\mu e_g}{2r_p+R_L}.$$

The average power output consumed in the load is

$$P_o=\frac{E_{max}I_{max}}{2}$$

$$=2R_L\left(\frac{\mu e_g}{2r_p+R_L}\right)^2.$$

As an example of push-pull class-A triodes, assume that the mu of each tube is 4 and that the plate resistance is 800 ohms; also assume that the peak signal voltage is 43 volts and that the effective load impedance (plate-to-plate) is 5,000 ohms. The power output is

$$P_o=2R_L\left(\frac{\mu e_g}{2r_p+R_L}\right)^2$$

$$=2\times 5{,}000\left(\frac{4\times 43}{2\times 800+5{,}000}\right)^2=6.8 \text{ watts.}$$

For class-A power pentodes, assume that the mu of each tube is 120 and that the plate resistance is 22,000 ohms; also assume that the peak-signal voltage is 16 volts and that the total load impedance is 5,000 ohms. The average power output is

$$P_o=2R_L\left(\frac{\mu e_g}{2r_p+R_L}\right)^2$$

$$=2\times 5{,}000\left(\frac{120\times 16}{2\times 22{,}000+5{,}000}\right)^2=15.3 \text{ watts.}$$

If one tube is removed from the circuit the voltage acting in the equivalent circuit is lowered and at the same time the

impedance in the plate load is reduced. Since the output impedance varies as the square of the number of turns, the removal of one tube leaves only one-half as many turns in the plate circuit. The load impedance is therefore $(\frac{1}{2})^2$, or $\frac{1}{4}$ of its former value, as indicated in figure 6–9, B.

The maximum plate current now becomes

$$I_{max}=\frac{\mu e_g}{r_p+\frac{R_L}{4}},$$

and the maximum voltage across the load becomes

$$E_{max}=I_{max}\frac{R_L}{4}$$

$$=\frac{\mu e_g}{r_p+\frac{R_L}{4}}\times\frac{R_L}{4}.$$

The average power output is then

$$P_o=\frac{I_{max}E_{max}}{2}$$

$$=\frac{\frac{\mu e_g}{r_p+\frac{R_L}{4}}\times\frac{\mu e_g}{r_p+\frac{R_L}{4}}\times\frac{R_L}{4}}{2}$$

$$=\frac{R_L}{8}\left(\frac{\mu e_g}{r_p+\frac{R_L}{4}}\right)^2.$$

If one of the pentodes of the preceding example is removed from the circuit, the power output becomes

$$P_o=\frac{5{,}000}{8}\left(\frac{120\times16}{22{,}000+\frac{5{,}000}{4}}\right)^2=4.25\text{ watts.}$$

This is a reduction in power of 72 percent.

The CLASS-B POWER AMPLIFIER is biased approximately to cutoff so that plate current flows in one tube during one half of the input cycle and in the other tube during the other half of the input cycle. Thus, one tube amplifies the positive half and the other the negative half of the input signal voltage. Both halves are combined in the secondary of the output transformer.

Since plate current flows in a given tube for only half of the input cycle and no appreciable current flows when no signal is applied, this type of amplifier has a higher efficiency than an amplifier operated class A or class AB. A class-B push-pull amplifier circuit is shown in figure 6–10, A.

The input-signal voltage curves shown in figure 6–10, B, make the grids alternately positive with respect to their associated cathodes on the positive peaks. When the grid voltage of triode A is positive maximum the plate voltage is minimum and the plate current is maximum. At this instant the reverse operation is taking place in triode B. The output signal voltages of the two tubes combine in series addition across the primary of the output transformer.

The plate voltage is alternately the sum and difference of the B-supply voltage and the induced voltage of one-half of the transformer primary.

The maximum plate current depends on the positive peak of the grid signal, and the minimum plate potential depends on the amount of voltage drop in the plate load. For best efficiency the minimum plate voltage should be as small as possible but not so small that the grid will draw excessive current.

The correct plate-to-plate load resistance in terms of I_{max}, E_{min}, and E_B of a single tube may be determined as follows: The current pulses from tubes A and B flow through only one-half of the primary of the output transformer; if they flowed through the entire primary, the amplitude would be only half as much, or $\frac{I_{max}}{2}$.

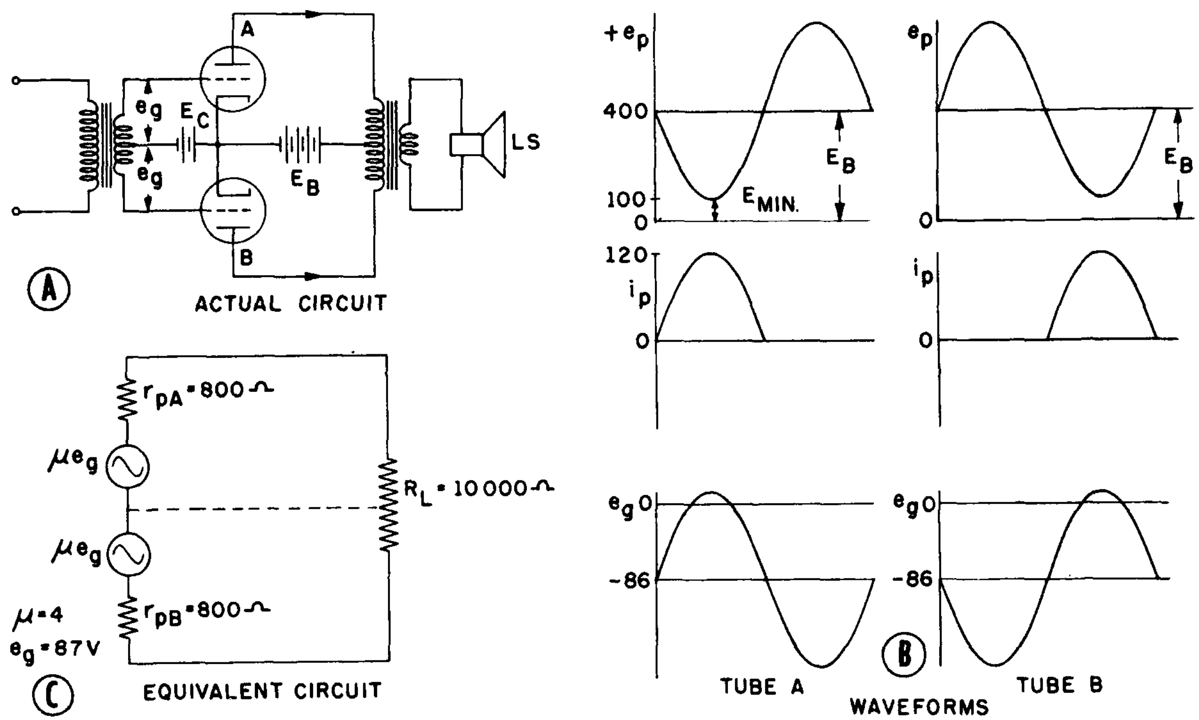


Figure 6-10.—Class-B operation.

The equivalent load resistance, R_L, acts in series with the triode plates and is equal to the impedance of the output transformer and its load as measured in terms of the primary. The voltage drop across R_L is $\frac{I_{max}}{2} R_L$. The alternating voltage drop between the plate and cathode of 1 tube is equal to one-half the voltage drop across R_L since the 2 tubes are in series. Thus,

$$e = \frac{1}{2} \frac{I_{max} R_L}{2} = \frac{I_{max} R_L}{4}.$$

Since the voltage across a single tube is equal to $E_B - E_{min}$, the equation may be rewritten as

$$\frac{I_{max} R_L}{4} = E_B - E_{min},$$

from which

$$R_L = 4 \frac{(E_B - E_{min})}{I_{max}}.$$

The average power output is equal to one-half the peak instantaneous power, or

$$P = \frac{I_{max}(E_B - E_{min})}{2}.$$

Using the values indicated on the curves of figure 6–10, B,

$$R_L = \frac{4(400-100)}{0.120} = 10{,}000 \text{ ohms};$$

and the average power output is

$$P = \frac{0.120(400-100)}{2} = 18 \text{ watts}.$$

The output power may also be determined by the use of the equivalent circuit shown in figure 6–10, C. If the maximum value of grid voltage, e_g, is assumed to 87 volts, the

maximum value of the signal current in the equivalent circuit will be

$$i=\frac{2\mu e_g}{2r_p+R_L}=\frac{2\times4\times87}{2\times800+10{,}000}=0.06 \text{ ampere.}$$

The maximum value of the signal voltage appearing across the load is,

$$e_L=iR_L$$
$$=0.06\times10{,}000=600 \text{ volts;}$$

and the average power output is

$$P_o=\frac{e_L i}{2}$$
$$=\frac{600\times0.06}{2}=18 \text{ watts.}$$

The output power of the class-B push-pull triode amplifier is about 2.6 times the output of the class-A triode amplifier given previously in this chapter.

Although approximately twice the signal voltage is applied to the grids of the class-B triodes as is applied to those of the class-A triodes, the output power is not 4 times as much because the load resistance of the class-B triode push-pull circuit is twice that of the class-A stage.

THE DECIBEL

Unit of Power Gain or Loss

The international transmission unit, the BEL, is a unit of gain equivalent to 10-to-1 ratio of power gain. Thus the gain in bels is simply the number of times that 10 is taken as a factor to equal the ratio of the output power of an amplifier to the input power. If, for example, the output power is 100 times the input, the ratio is 100 to 1, or 10^2 to 1. The gain is therefore 2 bels; and the gain in decibels (db) is 10 times 2, or 20 decibels.

The number of 10 factors contained in the ratio of the output power to the input power is the logarithm of the ratio to the base 10. The gain in decibels may therefore be expressed conveniently as

$$db = 10 \log_{10} \frac{P_2}{P_1}, \tag{6-4}$$

where P_2 and P_1 are respectively the output and input power in watts.

The human ear responds to ratio changes in intensity rather than to changes in absolute value. In other words, the ability of the human ear to detect changes in the intensity of sound is much greater at low levels of intensity than it is at high levels. A change in power level of 1 db is barely perceptible to the ear, and for this reason attenuators in audio systems are frequently calibrated in steps of 1 db.

Since the ear responds logarithmically to variations in sound-intensity levels, any practical system for measuring sound-intensity levels must necessarily vary logarithmically. The decibel system of measuring power levels is based on this concept.

Since the gains or losses in a system are expressed logarithmically, they are simply added or subtracted to determine the over-all gain or loss. For example, transmission lines introduce a loss in power, amplifier stages produce a gain, and attenuators introduce a loss. The final result is the algebraic sum of the various gains and losses. A db gain or loss is readily determined by the use of equation (6–4).

Current and Voltage Ratios

Primarily, the decibel is a unit to measure a power ratio. It can be used readily to compute current ratios as well, provided the resistances through which the currents flow are taken into account. The db gain or loss expressed in terms of the crrrents and resistances is determined as follows:

$$db = 20 \log_{10} \frac{I_2\sqrt{R_2}}{I_1\sqrt{R_1}}. \tag{6-5}$$

where I_2 and I_1 are respectively the output and input current in amperes, and R_2 and R_1 are respectively the output and input resistances in ohms. Thus, if the two currents and resistances are known, the db gain or loss can be determined by substitution in equation (6–5). If the resistances are equal they may be canceled out.

The same reasoning also applies to the voltage ratio provided the resistances across which the voltages are applied are properly considered. The equation for db gain or loss when voltages and resistances are employed directly is determined as follows:

$$\text{db}=20 \log_{10} \frac{E_2\sqrt{R_1}}{E_1\sqrt{R_2}}, \tag{6–6}$$

where E_2 and E_1 are respectively the output and input voltages, and R_2 and R_1 are respectively the output and input resistances in ohms.

If the voltages and resistances are known, the db gain or loss may be determined by direct substitution in equation (6–6). If the resistances are equal they may, of course, be canceled out.

Reference Levels

Considerable confusion has resulted from the use of various so-called zero-power reference levels. The term "zero reference level" is in itself somewhat confusing because it does not mean that no power is developed at that level. It means rather that the output level is referred to an arbitrary level designated as the reference, or zero, level; and as such it is perhaps one of the most convenient ways of expressing a power ratio. It is thus meaningless to say, for example, that a certain amplifier stage has a gain of 30 db unless reference is made to some established power level.

It is common practice in telephone work to consider 6 milliwatts as the reference power level. Other values

are also used in this and other fields, for example, 1, 10 and 12.5 milliwatts, depending upon which unit is most convenient under the circumstances.

The VOLTAGE gain or loss of microphones, transmission lines, and voltage amplifiers is also generally expressed in decibels. In general, transmission lines introduce a loss and voltage amplifiers produce a gain. A reference voltage level and the resistance (if it differs from the one being cómpared) across which the signal appears must be given in order that the gain or loss may have meaning.

The voltage output of a microphone may be expressed in terms of decibels below 1 volt per dyne per cm². In other words, 1 dyne acting on 1 square centimeter and producing an output of 1 volt is taken as the zero-decibel output level.

When the voltage gain of an amplifier stage is given in decibels the impedance must be given or they must be assumed to be equal. Thus if they are assumed to be equal, the gain in decibels may be expressed as

$$\text{db}=20 \log_{10} \frac{E_2}{E_1},$$

where E_2 is the output voltage and E_1 the input voltage.

When certain arbitrary power reference levels are used, the db gain or loss is given a special designation. One of these designations is the dbm, or the power level in decibels referred to 1 milliwatt, as

$$\text{dbm}=10 \log_{10} \frac{P}{0.001},$$

where P is the power in watts.

Typical power levels in dbm (0.001 watt in 600 ohms) are shown in figure 6–11 for various applications in audio systems.

The volume level of an electrical signal made up of speech, music, or other complex tones is measured by means of a specially calibrated voltmeter called a VOLUME INDICATOR. The volume levels registered on this indicator are expressed in volume units (VU). The number of units is numerically equal to the number of decibels above or below the reference volume level. Zero VU represents a power of 1 millwatt

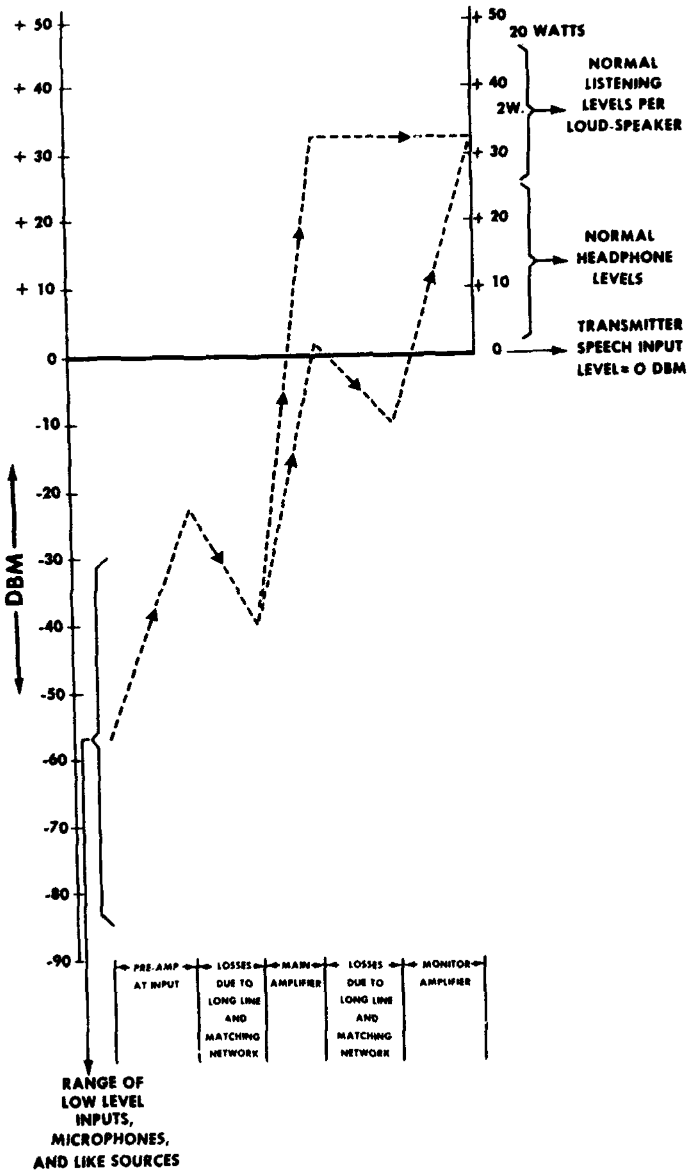


Figure 6-11.—Typical power levels in dbm (0.001 watt in 600 ohms) for various parts in audio systems.

dissipated in an arbitrary load resistance of 600 ohms (corresponding to a voltage of 0.7746 volts). Thus, when the VU meter is connected to a 600-ohm load, VU readings in decibels can be used as a direct measure of power above or below a 1-milliwatt reference level.

Use of Decibels

Some practical examples illustrating the use of decibels will perhaps clarify the foregoing paragraphs. (Logarithms to the base 10 are called COMMON LOGARITHMS. When the subscript following the logarithm is omitted, the base is understood to be 10. The following examples involve common logarithms.)

1. How many decibels correspond to a power ratio of 100?

$$\begin{aligned} \text{db} &= 10 \log \frac{P_2}{P_1} \\ &= 10 \log 100 \\ &= 10 \times 2 = 20. \end{aligned}$$

2. How many decibels correspond to a voltage ratio of 100 (assume equal resistances)?

$$\begin{aligned} \text{db} &= 20 \log \frac{E_2}{E_1} \\ &= 20 \log 100 \\ &= 20 \times 2 = 40. \end{aligned}$$

3. If an amplifier has a 20-db gain, what power ratio does this gain represent?

$$\frac{P_2}{P_1} = x$$

$$\text{db} = 10 \log x$$

$$20 = 10 \log x$$

$$\log x = 2$$

$$x = 100.$$

4. If an amplifier has a 30-db gain, what voltage ratio does this gain represent (assume equal resistances)?

$$\frac{E_2}{E_1}=x$$

$$\text{db}=20 \log x$$

$$30=20 \log x$$

$$\log x=1.5$$

$$x=31.6.$$

The voltage ratio is 31.6 to 1.

5. A certain microphone rated at −75 db is connected to a preamplifier through an attenuator rated at −10 db. The final audio amplifier is driven by the preamplifier and has a gain of 30 db. What must be the db gain of the preamplifier if the full output is to be achieved? (All db gains or losses have the same reference level.)

The total loss is 75+10 or 85 db and therefore the preamplifier must have a gain of 85 db to bring the losses to 0 db. From this point the main amplifier increases the gain 30 db above the common, or zero, reference level.

6. If the input to a certain loudspeaker is increased from 5,000 milliwatts to 6,000 milliwatts could the increase in volume level be readily detected by the human ear?

$$\begin{aligned}\text{db}&=10 \log \frac{P_2}{P_1}\\ &=10 \log \frac{6{,}000}{5{,}000}\\ &=10 \log 1.2\\ &=10\times 0.0792\\ &=0.792\end{aligned}$$

Since a change of 1 db is barely discernible, a change of 0.792 db would probably not be detected.

7. If 1 volt is applied across the 600-ohm input of a certain amplifier and 500 volts is developed across the 5,000-ohm output, what is the db power gain?

$$P_2=\frac{E_{out}^2}{R_{out}}=\frac{500^2}{5,000}=50 \text{ watts}$$

$$P_1=\frac{E_{in}^2}{R_{in}}=\frac{1}{600}=0.00166 \text{ watt}$$

$$\text{db}=10 \log \frac{P_2}{P_1}$$

$$=10 \log \frac{50}{0.00166}$$

$$=10 \log 30,000$$

$$=10\times 4.4770$$

$$=44.8.$$

8. If the noise level in a certain transmission line is 50 db down from the desired signal level of 10 mw, how much power is contained in the noise level?

$$\text{db}=10 \log \frac{P_2}{P_1}$$

where P_1 is the power in milliwatts contained in the noise level and P_2 is the power in milliwatts contained in the desired signal level.

Substituting,

$$50=10 \log \frac{10}{P_1}$$

$$5=\log \frac{10}{P_1}$$

$$10^5=\frac{10}{P_1}$$

$$P_1=\frac{10}{10^5}=10^{-4}=0.0001 \text{ mw.}$$

QUIZ

1. What is the primary function of a power amplifier?
2. Give three general characteristics of audio power amplifiers.
3. In a class-A power amplifier, what characteristics of the i_p-e_g curve limits the minimum value of i_p?
4. For maximum undistorted power output in a class-A triode power amplifier, what is the relative value of load impedance with respect to the plate resistance of the triode?
5. Why is the efficiency of a class-A amplifier low?
6. State the equation of the load line for a power amplifier in terms of the plate current, plate load resistance, and plate supply voltage.
7. What is the theoretical value of plate voltage as indicated on the load line in figure 6–1 for the condition of maximum plate current?
8. When no plate current flows, what is the value of plate voltage as indicated by the load line in figure 6–1?
9. In figure 6–1, what is the value of plate current when no signal is applied?
10. If the maximum value of the second harmonic current is 0.8 ma and the maximum value of the fundamental is 20 ma, what is the percentage of distortion due to the second harmonic?
11. Express the equation of the primary-to-secondary turns ratio of a transformer in terms of the primary-to-secondary matching impedances.
12. What causes the reduced high-frequency response of an output transformer?
13. What characteristics must an output transformer have to extend the flat portion of the frequency-response curve of the output transformer into both the low- and the high-frequency regions?
14. Name four advantages of using push-pull amplifiers in the output sta of an a-f amplifier?
15. What is the phase relation between the two second-harmonic components in figure 6–8?
16. If one tube is removed from a typical class-A push-pull pentode power amplifier, what is the percentage of power reduction?
17. Why is a class-B push-pull amplifier more efficient than a class-A or class-AB amplifier?
18. What is the relative output power of a triode class-B push-pull amplifier compared with that of a class-A amplifier?
19. Why is the output power of a triode class-B push-pull circuit less than 4 times the output power of a similar class-A triode if the former has twice as much input voltage?

20. If the output power of an amplifier is 1,000 times the input power, what is the gain in decibels?
21. Why should any practical system for measuring sound-intensity levels vary logarithmically?
22. What is the common term for the arbitrary reference level in expressing power ratios?
23. An amplifier has a gain of 30 dbm. What is the power output of the amplifier?

CHAPTER 7

OSCILLATORS

INDUCTANCE-CAPACITANCE OSCILLATORS

Introduction

The primary function of an oscillator is to generate a given frequency and to maintain that frequency within certain limits. To that end inductance-capacitance oscillators depend for their operation on the resonant interchange of energy between a capacitor and an inductor, with an electron-tube amplifier supplying pulses of energy of the proper phase and magnitude to maintain the oscillations. Resonant circuits (tuned circuits) are treated in chapter 1, and electron-tube amplifiers are treated in chapters 4 and 5.

In addition to their use as amplifiers, electron tubes are used as oscillators for the generation of alternating voltages. When thus used as oscillators, electron tubes are essentially converters that change d-c electrical energy from the plate power supply into a-c electrical energy in the output circuit. To accomplish this energy conversion, the amplifying ability of the electron tube is used in such a manner as to generate sustained oscillations.

Two conditions are necessary if sustained oscillations are to be produced. FIRST, the feedback voltage from the plate circuit must be in phase with the original excitation voltage on the grid—that is, the feedback must be positive, or regenerative. SECOND, the amount of energy fed back to the grid

circuits must be sufficient to compensate for the energy losses in the grid circuit.

Feedback may be accomplished by inductive, capacitive, or resistive coupling between the plate and the grid circuit. Various circuits have been developed to produce feedback of the proper phase and amount. Each of these circuits has certain characteristics that make its use advantageous under given circumstances.

If the proper values of inductance and capacitance are used, tuned-circuit oscillators may be designed to generate frequencies from the low frequencies in the audio range to the very high radio frequencies The upper frequency limit is determined in general by the distributed inductance and capacitance of the circuit components and the interelectrode capacitance of the tubes.

The electron tube itself is not an oscillator. The oscillations actually take place in the tuned circuit, a part of which may be composed of the interelectrode capacitances of the electron tube and the distributed capacitances and inductances of the circuit. The electron tube functions primarily as an electrical valve that amplifies and automatically releases to the grid circuit the proper amount of energy to maintain oscillation.

A basic oscillator showing the circuits necessary for its operation is shown in figure 7–1.

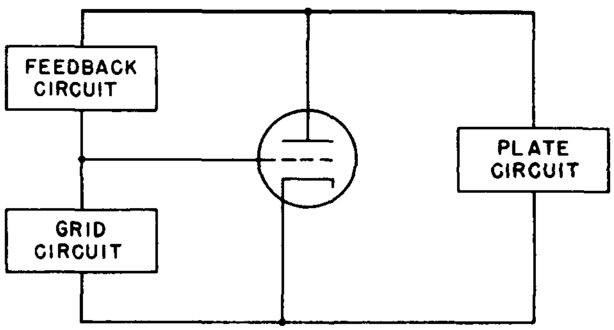


Figure 7–1.—Basic oscillator circuit.

Tickler-Feedback Oscillator

One of the simplest types of oscillator circuits is that employing tickler feedback, as shown in figure 7–2. Feed-

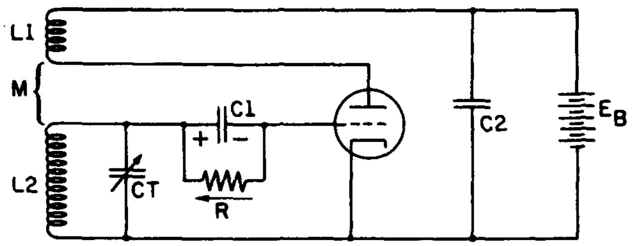


Figure 7–2.—Tickler-feedback oscillator circuit.

back voltage of the proper phase from the plate to the grid circuit is accomplished by mutual inductive coupling between the oscillator tank coil, *L2*, and the tickler feedback coil, *L*1. The amount of feedback voltage is determined by the amount of flux from *L*1 that links *L*2. Thus the feedback voltage is varied by moving *L*1 with respect to *L*2 or by changing the setting of a variable resistor that is sometimes shunted across *L*1.

The frequency-determining part of the oscillator is the tank circuit, *L2CT.* The coil and tuning capacitor interchange energy at the resonant frequency rate and the excitation voltage developed across *CT* is applied to the grid in series with the grid-leak bias across *RC*1.

Grid current flowing through *R* establishes negative bias on the grid. Capacitor *C*1 charges up to the peak voltage across *R* and holds this voltage between r-f pulses because of the relatively long time constant of *RC*1 as compared with the time for each r-f cycle. A steady bias voltage is therefore developed when the oscillator is functioning properly. Thus, a test for proper operation is a measurement of the d-c voltage across the grid resistor. To avoid blocking the oscillations, a high-resistance type of meter, such as the electron-tube voltmeter, should be used for this measurement.

Series-Fed Hartley Oscillator

Figure 7-3 shows the circuit of a series-fed Hartley oscillator and the curves of grid voltage, plate current, and plate voltage for class-C operation. The plate circuit in figure 7-3, A, is shown in heavy lines. Thus, *L*1 is a part of the tuned circuit made up of *L*1, *L*2, and *CT*; it also serves to couple energy from the plate circuit back into the tuned grid circuit by means of the mutual inductance between *L*1 and *L*2. Capacitor *C*1 blocks the d-c component of grid

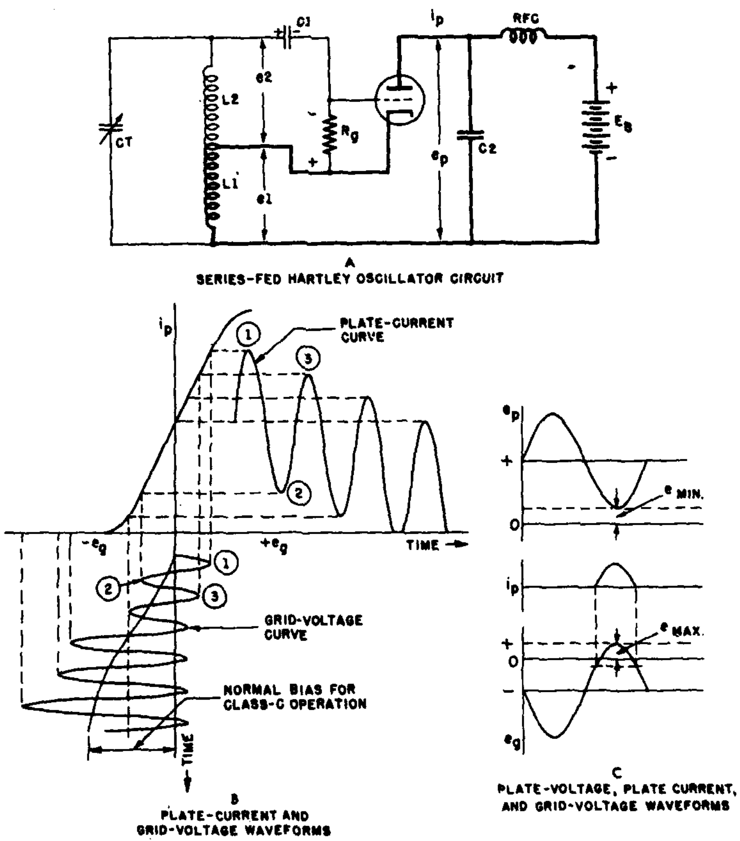


Figure 7-3.—Analysis of a series-fed Hartley oscillator.

current from $L2$ and, together with R_g, provides the necessary operating bias. Capacitor $C2$ and the radio-frequency choke (RFC) keep the alternating component in the plate circuit out of the B-supply.

The B-supply in this example is returned to the resonant tank coil, $L1$. The tuned circuit therefore contains a d-c component of plate current in addition to the a-c signal component. Thus the B-supply is connected in series with the plate and tank coil $L1$. This is a SERIES-FED connection.

The series-fed Hartley oscillator operates as follows (fig. 7-3):

1. When the tube warms up, plate current starts to flow. Because the grid is positioned in the electric field (between the cathode and plate) at a point that is positive with respect to the cathode, a small positive voltage exists on the grid.
2. The increase in plate current through $L1$ is accompanied by an expanding magnetic field around $L1$, which induces voltage e_2 in $L2$. The polarity of e_2 makes the grid more positive with respect to the cathode, and plate current increases to saturation (point 1, fig. 7-3, B). Capacitor CT is charging. Grid current flows as $C1$ acquires a small charge with the polarity indicated in the figure. The grid voltage during this time is e_2 minus the drop across $C1$.
3. Plate current stops rising at saturation and the field about $L1$ stops expanding. The induced voltage, e_2, falls to zero.
4. The positive grid voltage (e_2 minus the drop across $C1$) decreases and the plate current decreases. Capacitor CT begins to discharge.
5. The field about $L1$ collapses and induces a voltage, e_2, in $L2$ of opposite polarity to e_2 when the field was expanding. Hence, the grid voltage is negative with respect to the cathode. Plate current decreases further.
6. The induced voltage, e_2, aids in the discharge of CT and $C1$. CT discharges fully and begins to charge in

the opposite direction (its polarity reverses). However, $C1$ cannot discharge rapidly because of the long time constant, R_gC1. Grid voltage swings to a maximum negative condition (point 2 on the grid-voltage curve) and $C1$ discharges slowly down through R_g. Grid current does not flow during this part of the cycle and the grid bias voltage is e_2 plus the drop across $C1$.

7. Plate current ceases to fall at point 2 on the plate-current curve corresponding to point 2 on the grid-voltage curve. The field about $L1$ stops changing and e_2 falls to zero. CT begins to discharge.
8. The grid bias voltage swings in a positive direction again, and plate current begins to rise. The expanding field about $L1$ again induces voltage e_2 in $L2$, making the grid voltage more positive with respect to the cathode. Current flows from cathode to grid into the left-hand plate and out of the right-hand plate of $C1$, causing $C1$ to acquire a small additional charge.
9. Plate current rises to saturation (point 3).
10. The cycle continues to repeat.
11. On each subsequent cycle the bias voltage builds up across $C1$ and R_g until it reaches a steady value, as shown in figure 7–3, B.
12. Normal bias indicates class-C operation. The flywheel effect (interchange of energy between coil and capacitor) of the resonant tank maintains the oscillations during the time that the plate current is zero and no energy is being supplied to the oscillator circuit.

The time constant of R_gC1 should be relatively long compared with the time for one oscillator cycle. If it is too short, the bias will be lost, and if it is too long the oscillator will be blocked periodically because of excessive accumulated bias. Self-bias makes the oscillator self-starting when the filament is energized and the plate voltage is applied.

Loading the oscillator may be accomplished by placing a relatively high resistance in shunt with the tank circuit or a low resistance in series with it. In either case the effec-

tive Q is reduced and the amplitude of the generated voltage is lowered. If the load is coupled inductively to the oscillator, the point of coupling should be near r-f ground in order to reduce any harmonic content in the output. Loading the oscillator may detune it. Buffer stages are frequently inserted between the oscillator and its load to isolate the oscillator and prevent frequency changes.

The resonant frequency of the tuned tank circuit is essentially the frequency of the oscillator. Expressed in terms of the resonant tank inductance and capacitance, the resonant frequency is

$$f_o=\frac{159}{\sqrt{LC}},$$

where f_o is the resonant frequency in megacycles, L the inductance in microhenrys, and C the capacitance in micromicrofarads.

For example, if the inductance of the tank circuit is 100 microhenrys and the capacitance is 100 micromicrofarads, the resonant frequency of the oscillator is

$$f_o=\frac{159}{\sqrt{100\times 100}}=1.59 \text{ megacycles.}$$

Shunt-Fed Hartley Oscillator

The shunt-fed Hartley oscillator (fig. 7–4) differs from the series-fed Hartley oscillator (fig. 7–3) in that direct current

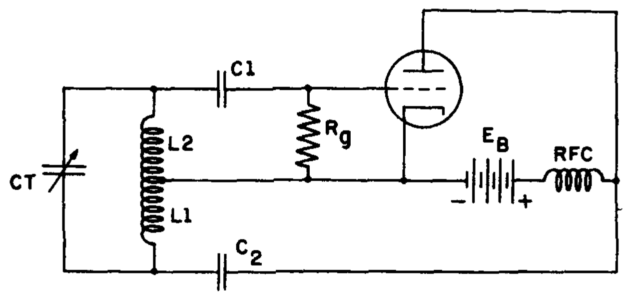


Figure 7–4.—Shunt-fed Hartley oscillator circuit.

does not flow through the tank circuit. Figure 7–4 indicates that the B-supply is connected in shunt with the triode plate and the portion of the tank circuit that includes $L1$. The d-c component of plate current is kept out of the $L1$ tank circuit by the blocking capacitor, $C2$, and the a-c component is kept out of the plate power supply by the radio-frequency choke coil.

An advantage of shunt feed is that the high-voltage B-supply is isolated from the tuned circuit.

Colpitts Oscillator

The Colpitts oscillator shown in figure 7–5 is similar to

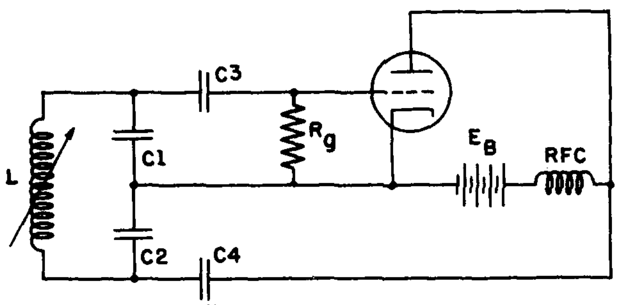


Figure 7–5.—Colpitts oscillator circuit.

the shunt-fed Hartley oscillator with the exception that the Colpitts oscillator uses a split tank capacitor as a part of the feedback circuit instead of a split tank inductor. The frequency of the oscillations is determined by the values of L, $C1$, and $C2$; and the grid excitation voltage appears across $C1$ instead of across $L2$ as in the shunt-fed Hartley oscillator. $C3$ and $C4$ perform the same function in this circuit as $C1$ and $C2$ in the shunt-fed Hartley circuit.

Ultraudion Oscillator

The ultraudion oscillator (fig. 7–6), frequently employed at ultrahigh frequencies, is similar to the Colpitts oscillator. The grid-to-cathode and plate-to-cathode interelectrode

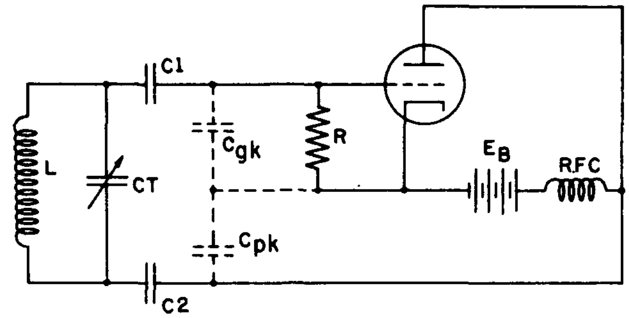


Figure 7–6.—Ultraudion oscillator circuit.

capacitances that make its operation similar to that of the Colpitts oscillator are indicated by dotted lines in the figure. Parallel feed is employed, and the radio-frequency choke prevents the a-c component of the plate voltage from entering the B-supply. Capacitor *C*2 provides a low-reactance path for r-f current and blocks direct current from the tank.

The voltage drop across C_{gk} is appreciable at the frequency employed and provides the grid excitation. Bias voltage is developed by the flow of grid current through *R*. The total tank capacitance is made up of *CT* in parallel with the series combination of *C*1, C_{gk}, C_{pk}, and *C*2. Capacitors *C*1 and *C*2 are relatively large so that they will offer negligible reactance to r-f currents.

Tuned-Plate Tuned-Grid Oscillator

The tuned-plate tuned-grid (TPTG) oscillator utilizes a tuned circuit in both the plate and the grid circuits, as shown in figure 7–7, A. This type of oscillator may be employed in a wide range of frequencies from the low frequencies to ultrahigh frequencies. However, because of reduced feedback between plate and grid at low frequencies, the TPTG oscillator is not so satisfactory at low frequencies as some of the circuits that have already been considered.

The feedback necessary to sustain oscillations is coupled from the plate circuit to the grid circuit by means of the interelectrode capacitance between the plate and the grid.

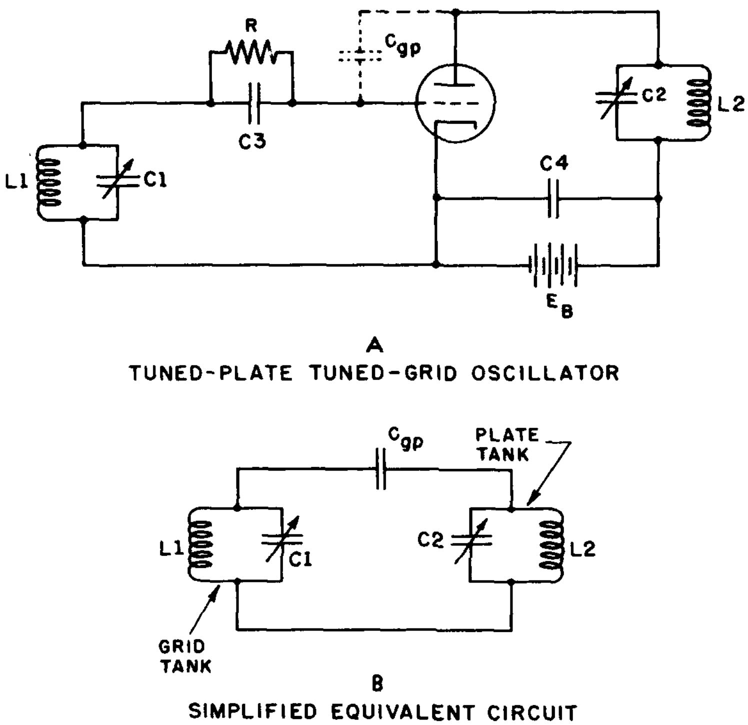


Figure 7-7.—Tuned-plate tuned-grid oscillator and equivalent circuits.

Self-bias is established by means of R and $C3$. Capacitor $C4$ is the r-f bypass around the B-supply.

The simplified equivalent circuit of figure 7-7, B, represents the feedback circuit of the oscillator as a series circuit containing the grid and plate tuned circuits in series with the interelectrode capacitance of the tube. At the frequency of operation of the oscillator, the plate tank presents inductive reactance to the equivalent series circuit which includes the interelectrode capacitance between plate and grid. In this respect the circuit is analogous to a series L-R-C circuit in which the inductive reactance neutralizes the capacitive reactance.

The output frequency of the oscillator is not quite equal to the natural resonant frequency of the tuned grid circuit or the tuned plate circuit. The output frequency is slightly less than the lower of the two tuned circuits. The tuned circuit having the lower frequency determines the output frequency of the oscillator.

Push-Pull Oscillator

In order to obtain a power output larger than is possible with a single tube, an additional tube may be added in push-pull. As in push-pull amplifiers, the harmonic content of the output is reduced. The frequency stability of the push-pull oscillator is increased over that of single-ended types. The effect of interelectrode capacitance is reduced and the frequency range is extended. Push-pull oscillators are used generally at high and ultrahigh frequencies.

The push-pull oscillator (fig. 7–8) utilizes the interelectrode

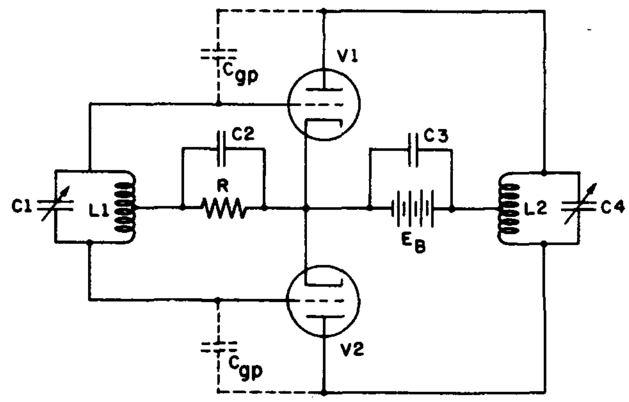


Figure 7–8.—Push-pull oscillator circuit.

capacitance of each tube to feed back to the grid tank sufficient voltage to sustain oscillations.

When the oscillator is first energized it is improbable that the two tubes will be operating under exactly the same conditions. One tube therefore conducts more current than the other, and the voltages fed back to the two grids are

unequal. An initial surge of energy is thereby introduced into the grid tank circuit and oscillations are set up because of the interchange of energy in the magnetic and electrostatic fields about $L1$ and $C1$, respectively. First one tube and then the other conducts; and in conducting, each tube feeds energy back to the tank circuit at the proper time to cause the voltage across the tank circuit to increase in amplitude.

The oscillations continue to increase in amplitude until the energy dissipated in the grid tank is equal to the energy supplied to it. The maximum amplitude of oscillation is called the SATURATION AMPLITUDE because the tubes are driven into the plate-current saturation regions of their characteristic curves.

Electron-Coupled Oscillator

If an oscillator is loaded appreciably, particularly when the load varies, the frequency stability of the oscillator is greatly reduced. Load changes in the power stages following the oscillator stage, however, have slight effect on the frequency stability of the oscillator.

The electron-coupled oscillator shown in figure 7-9 com-

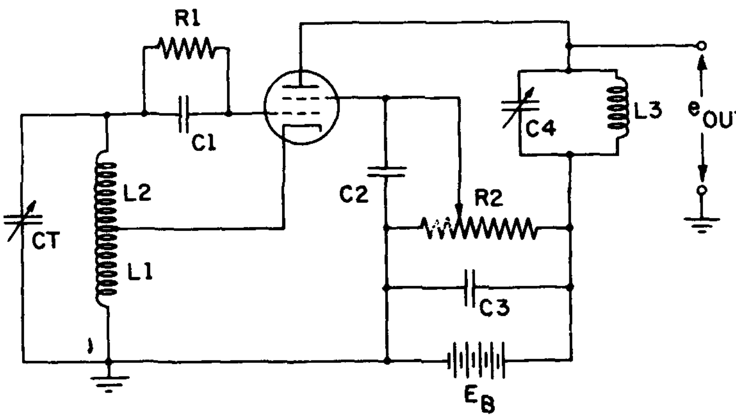


Figure 7-9.—Electron-coupled oscillator circuit.

bines the functions of an oscillator and a power amplifier. The control grid, tank circuit (*CT*, *L*1, and *L*2), cathode, and screen grid form a series-fed Hartley oscillator with the screen grid serving as the plate. Capacitor *C*2 places the screen at zero r-f potential and, like *C*3, bypasses the plate supply.

The output tuned circuit, *C*4*L*3, is in the plate circuit. The electron stream is the only coupling medium between the grid tank and the plate tank—hence the name ELECTRON-COUPLED OSCILLATOR. The two tank circuits are isolated by the screen grid, which is at r-f ground potential. This type of oscillator is relatively stable. Load variations have slight effect on the frequency of the oscillations.

An increase in screen voltage decreases the frequency of the oscillator, while an increase in plate voltage increases the frequency. If the screen and plate voltages are derived from the same power supply and the supply voltage increases, the tendency of the plate to increase the frequency and the screen to decrease the frequency cancel each other and the frequency remains unchanged. Conversely, a decrease in supply voltage has a similar effect. The tendency of the plate to decrease the frequency and the screen to increase the frequency again cancel each other. Potentiometer *R*2 is adjusted for optimum screen voltage after which no further adjustment should be necessary. Frequency stability is best when the ratio of the plate-to-screen voltage is about 3 to 1.

Negative-Resistance Oscillator

A negative-resistance effect exists in a circuit if an increase in voltage across the circuit is accompanied by a decrease in current through it. This negative resistance effect was noted in the discussion of screen-grid tubes. Thus, an increase in plate voltage of a tetrode is accompanied by a decrease in plate current, provided the plate voltage does not exceed that of the screen.

Another concept of negative resistance is applied to regenerative feedback in an oscillator. In supplying the

input power to the osciliator, regenerative feedback introduces a negative resistance. In this sense negative resistance represents a source of power that supplies the losses in the oscillator circuit.

In the various feedback oscillator circuits that have been considered, enough power was available to overcome the circuit losses and to supply the necessary external load. If all of the circuit and load losses were represented by an equivalent resistance in the plate circuit, the amplifier action of the tube might be considered as presenting a negative resistance to the tuned circuit. The power supplied by the negative resistance must be equal to the power consumed by the positive resistance.

The use of negative transconductance is a common method of obtaining negative resistance, as in the transitron oscillator. Plate resistance, r_p, may be expressed as $\frac{\mu}{g_m}$. If g_m is negative, r_p is negative.

TRANSITRON OSCILLATOR.—The circuit of an elementary transitron oscillator utilizing negative transconductance is shown in figure 7-10. As indicated in the figure, the plate potential is lower than the screen-grid potential. Also, the suppressor grid is biased slightly negative with respect to

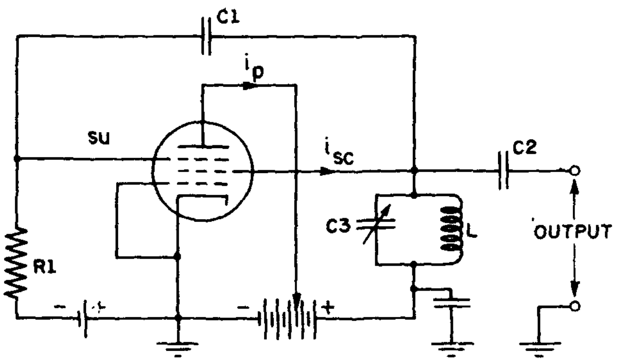


Figure 7-10.—Elementary transitron oscillator circuit.

the cathode. Under these conditions the plate current, i_p, is greatly reduced because the plate voltage has less influence on the electrons leaving the cathode, and many of the electrons that do succeed in passing through the screen grid are repelled back to it by the negative charge on the suppressor grid. The plate and screen currents may be approximately equal.

Negative resistance is produced by the action of the negative bias applied to the suppressor grid. Electrons that would normally pass through the suppressor to the plate are repelled back to the screen grid. If the negative bias on the suppressor is reduced, fewer electrons are repelled by the suppressor back to the screen, and screen current decreases. In figure 7–10 the change in suppressor potential is accomplished through coupling capacitor $C1$. Any change in the screen tank voltage is passed through $C1$ to the suppressor grid. The suppressor potential is more effective in controlling screen current than is the screen potential. Hence, a given decrease in suppressor bias caused by a corresponding increase in screen potential results in a decrease in screen current as more electrons flow to the plate.

Thus an INCREASE in screen potential (and an equal decrease in suppressor bias via $C1$) causes a DECREASE in screen current, producing NEGATIVE RESISTANCE. By the use of negative resistance, oscillations may be maintained in the tuned circuit, $C3L$, which is connected between the suppressor grid and the B-supply. When the circuit is oscillating, the alternating component is applied simultaneously to the screen and suppressor grids. When the voltage is more positive, i_{sc} is reduced; when it is less positive, i_{sc} is increased.

The characteristic curve of screen current vs screen voltage, when the screen and suppressor grids are coupled by means of a capacitor, is shown in figure 7–11. The screen-grid potential is adjusted so that it operates on the negative slope of the curve.

Figure 7–12 shows the variations of i_p and i_{sc} with changes

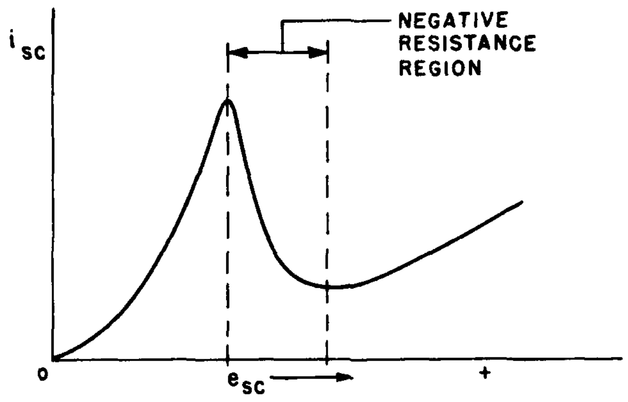


Figure 7–11.—Screen-current vs screen-voltage characteristic of a pentode with screen-suppressor coupling.

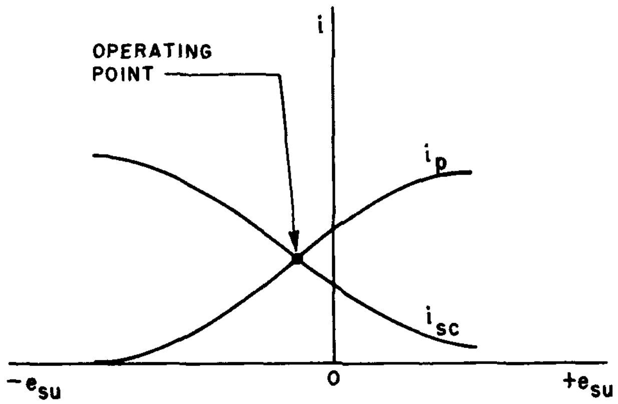


Figure 7–12.—Plate and screen currents as functions of suppressor voltage.

in voltage on the suppressor grid, for the conditions under which the tube operates. The negative slope of the i_{sc} curve indicates that the transconductance between the suppressor-grid voltage and the screen-grid current is negative. Because the reactance of $C1$ is negligible at the oscillator frequency,

the alternating component on both grids is of the same instantaneous polarity. Thus, the negative transconductance of the tube becomes a negative resistance between the screen grid and the cathode. An increase in screen voltage causes a corresponding increase in suppressor voltage and hence a decrease in screen current.

The losses in the tuned circuit must be offset by energy supplied through the tube if oscillations are to be sustained. The equivalent circuit of the transitron oscillator is shown in figure 7-13. R_N is the negative resistance presented by the

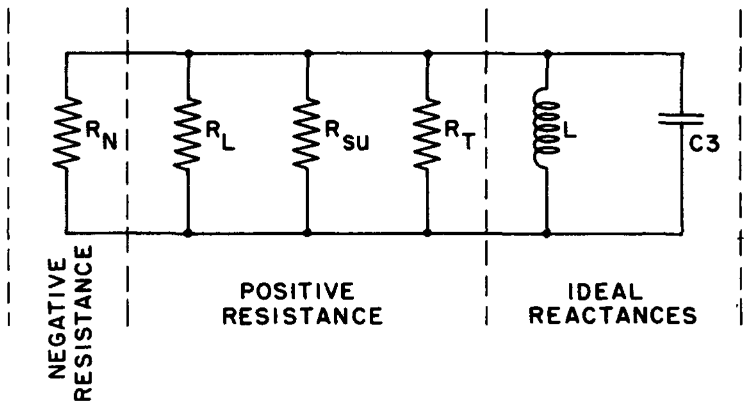


Figure 7-13.—Equivalent circuit of a transitron oscillator.

tube to the tuned circuit; R_L represents the load connected across the output terminals; R_{su} represents the suppressor-grid resistance; and R_T represents the tuned-circuit losses. L and $C3$ are the inductance and capacitance, respectively, of the tuned circuit.

In order to produce oscillations of constant amplitude, the power supplied by R_N must equal the power consumed by the three positive resistances. Therefore, the current through R_N must be equal to the combined currents through the three positive resistances. If less current flows through R_N than through the positive resistances, the oscillations will die out,

and if more current flows through R_N than through the positive resistances, the oscillations will increase in amplitude. When the current through R_N is just sufficient to sustain oscillations, the equivalent circuit reduces to one composed of ideal reactances—that is, to the simple *L-C* combination making up the tuned circuit. The frequency of oscillations in the tuned circuit is then expressed as

$$f_o=\frac{1}{2\pi\sqrt{LC}}.$$

If the transitron oscillator is to be used to produce CONTINUOUS oscillations, the circuit is adjusted so that R_N is smaller than the value required to sustain oscillations. The transient oscillations that occur when the B-supply switch is closed are amplified. The suppressor-grid voltage swings less negative and more negative about the operating point (the point of greatest slope), which is biased slightly negative with respect to the cathode. (See fig. 7–12.) The operating range on the i_{sc}-e_{su} curve extends to the right and to the left of the operating point, and thus the average slope of the part used is decreased. This decrease, in effect, increases the value of the negative resistance, and the amplitude of the oscillations increases until the value of R_N is sufficient to maintain a constant amplitude,

Crystal-Controlled Oscillator

CRYSTAL CHARACTERISTICS.—Some crystalline substances such as Rochelle salt, quartz, tourmaline, and even cane sugar, have the property of changing their shape when an emf is impressed upon the crystal. The action is also reversible—that is, if the crystal is subjected to a mechanical strain, an emf will be produced across the surfaces of the crystal. A strain in one direction produces a certain polarity between the surfaces; if the direction of the strain is reversed, the polarity of the emf will also be reversed. This interrelation between mechanical and electrical stress in crystals is called the PIEZOELECTRIC EFFECT.

The magnitude of the response obtained from the crystal depends on the type of crystal used, the way it is cut, and the manner in which the emf is impressed. Roohelle salt is perhaps the most active, but it is also affected to a large extent by heating, aging, mechanical shock, and moisture. Quartz, while less active, is more rugged than Rochelle salt.

Another and more important advantage of the quartz crystal is its inherently higher Q. This results from the low damping, which in turn is due to the hardness of the crystal and its low internal friction when vibrating. Values of Q above 25,000 are possible; and if air damping is reduced sufficiently, Q's of the order of several hundred thousand are possible.

Since a high Q is necessary for frequency stability, quartz-crystal oscillators are widely used.

Quarts crystals used in oscillator circuits must be cut and ground to accurate dimensions. For example, the dimensions for a typical quartz crystal resonant at 1,000 kc would be approximately $1 \times 1 \times 0.1125$ inch. All crystals, however, are not in the shape of a rectangular plate such as this. For use at the higher frequencies some of the crystal elements are disk shaped, similar to a coin. For precise test work, crystals may be cut in the form of a flat ring.

Electrical contact with the crystal is made by means of a crystal holder consisting of two metal plates, between which the crystal is placed, and a spring device that places mechanical pressure on the plates.

Crystals are classified according to the way they are cut from the original quartz crystal. Figure 7–14 shows the approximate form of a raw crystal, a cross section of the crystal, and a section of the crystal in which three of the possible cuts are indicated.

The Z, or optical, axis is not important electrically because no piezoelectric effect is produced by the application of electrical stresses in this direction. The X axes drawn through the corners of the hexagon (fig. 7–14, B) are called the ELECTRICAL AXES, and the Y axes drawn perpendicular

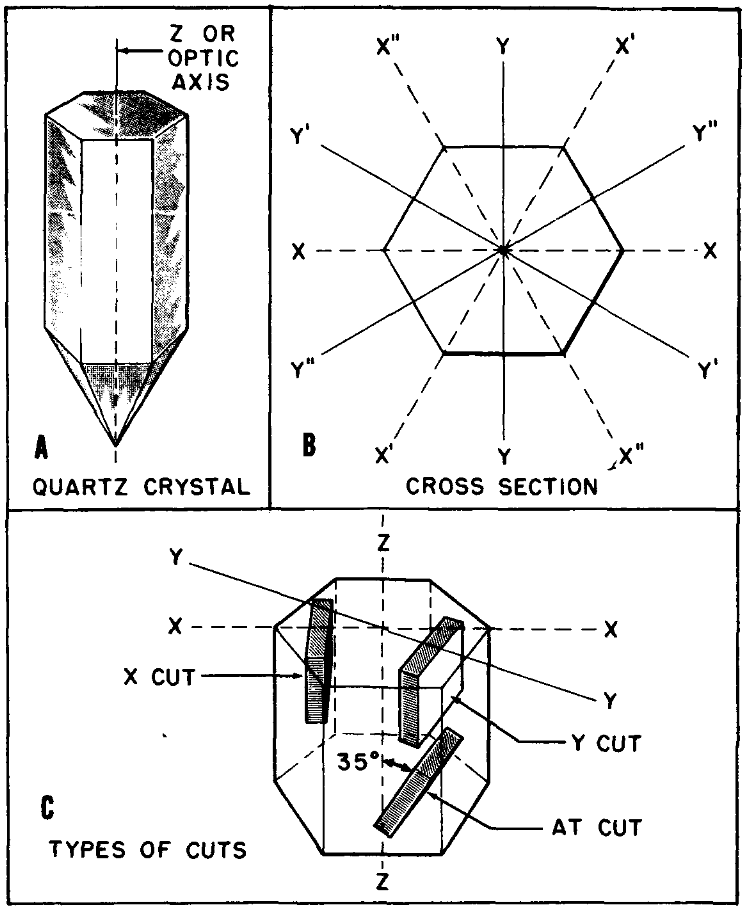


Figure 7–14.—Quartz crystal.

to the faces of the hexagon are called the MECHANICAL AXES. Designation of the axes as electrical or mechanical, aids in defining the type of cut. For example, in figure 7–14, C, the *X*-cut crystal is shown cut perpendicular to one of the *X* axes. During one-half of the r-f voltage cycle impressed across the flat surfaces of this *X*-cut crystal it will expand

along the axis perpendicular to its flat surfaces, and during the other half cycle it will contract. Likewise, the *Y*-cut crystal is cut perpendicular to one of the *Y* axes. The *AT* cut is made at approximately a 35° angle with the *Z* axis. Each of the other cuts such as the *BF*, *CF*, *DT*, and *GT* cuts, has distinctive characteristics.

Temperature has different effects on the various types of cuts. The *X*-cut crystal has a negative temperature coefficient—that is, as the temperature increases, the frequency decreases. The *Y*-cut crystal has a positive temperature coefficient—as the temperature increases, the frequency increases. Both of these effects are undesirable, and therefore these two types of cuts have been replaced by others. The *AT*-cut shown in figure 7–14, C, has a temperature coefficient that is very nearly zero—that is, its frequency changes only slightly with changes in temperature. The *GT*-cut (not shown) has the lowest temperature coefficient of any of the cuts.

Where the frequency of the oscillator must be maintained within a few cycles of the assigned frequency, as in standard broadcast transmitters, the crystal is placed in a temperature-controlled chamber. Changes in frequency due to changes in temperature are thus maintained at a minimum value, consistent with the ability of the thermostats to keep the temperature constant.

When a crystal starts vibrating at its resonant frequency, it requires only a small force operating at the same frequency to obtain vibrations of a large amplitude. When an alternating voltage is applied to a crystal that has the same mechanical frequency as the applied voltage, it vibrates and only a small applied voltage is needed to keep it vibrating. In turn, the crystal generates a relatively large voltage at its resonant frequency.

If a crystal is placed between the grid and cathode of an electron tube, and a small amount of energy is fed back from the plate circuit of the crystal to keep it oscillating, the circuit will act as an oscillator. The natural frequency of a crystal is critical, and if the frequency fed back is slightly

higher or lower than this value the crystal will stop vibrating. Thus, the frequency of the crystal-controlled oscillator must be the same as that of the natural frequency of the crystal.

CIRCUIT OPERATION.—A crystal-controlled oscillator employing a triode tube is shown in figure 7–15. The equivalent

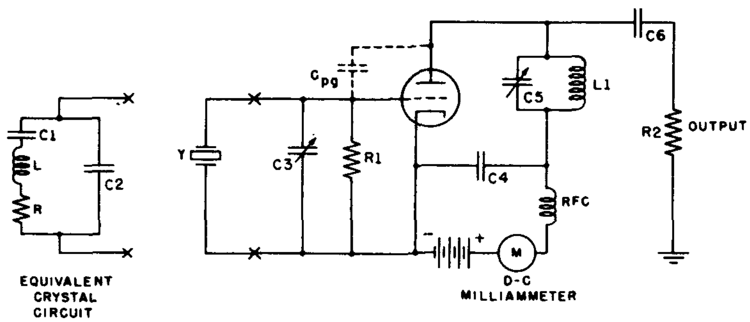


Figure 7–15.—Crystal-controlled oscillator circuit using a triode tube.

electrical circuit of crystal *X* is also shown. In the equivalent circuit, the inductance, *L*, represents the electrical equivalent of the crystal mass that is effective in causing mechanical vibration; *R* is the electrical equivalent of internal resistance due to friction; *C*2 is the capacity effect of the metal crystal holders; and *C*1 is the reciprocal of the crystal stiffness—that is, COMPLIANCE, which is the equivalent of capacitance in the electrical system.

Feedback takes place through the plate-to-grid capacitance within the electron tube, and the bias voltage is established across *R*1. Capacitor *C*3 has a small capacitance and is shunted across the crystal in order to obtain a fine adjustment of the operating frequency by changing the capacitance of the equivalent electrical circuit. The output voltage appears across *R*2.

Oscillations occur at the resonant frequency of the crystal, and the plate circuit is tuned to a slightly higher frequency by decreasing the capacity of *C*5 so that the plate tank presents inductive reactance to the crystal grid circuit.

ADJUSTING A CRYSTAL-CONTROLLED OSCILLATOR.—In ad-

justing a crystal oscillator, the factor of stable operation must be considered. In figure 7–15, a d-c milliammeter is connected in series with the B+ lead to the plate tank circuit, and $C5$ is changed from a low to a high value of capacitance (tuned to a lower frequency). The plate current will slowly decrease to a MINIMUM at the exact resonant frequency of the crystal oscillator, as shown at point C in figure 7–16. At this point the output of the plate tank

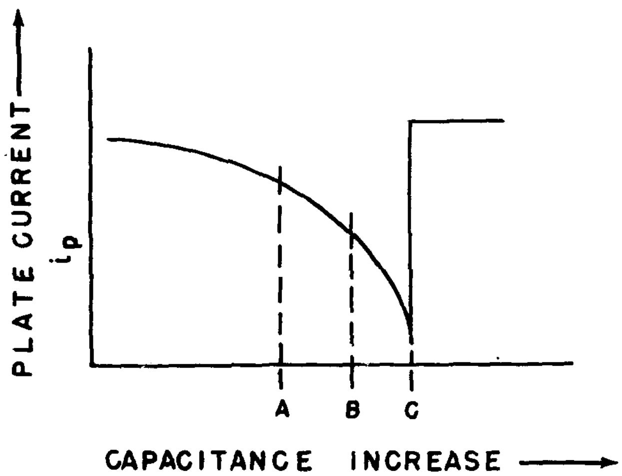


Figure 7–16.—Crystal-oscillator plate-current tuning curve.

circuit is maximum (minimum direct current indicates maximum a-c output). Just to the right of point C the plate current suddenly increases to its maximum value, and oscillations cease.

The crystal frequency is slightly below the natural resonant frequency of the plate tank, and the tank, therefore, looks like an inductor. The feedback circuit is a series connection of the plate tank, grid crystal, and plate-to-grid capacitance of the triode. As long as the feedback is positive the oscillator operates at the crystal frequency. As the capacitance is increased beyond point C, the tank suddenly looks like a

capacitor instead of an inductor and the feedback becomes negative instead of positive. Oscillations therefore cease, and plate current rises rapidly.

In order to stabilize the operation, the capacitance is decreased (the resonant frequency is increased) to a value somewhere between A and C—for example, at B. The output is thereby reduced, but the operation is much more stable and slight changes in loading will not cause the oscillator to cease functioning.

The strength with which the crystal vibrates at its resonant frequency depends on the voltage fed back to it. The feedback is controlled by adjustment of the tuning of the plate tank circuit. If the feedback is too great, the vibrations may have sufficient magnitude to crack the crystal. The use of tetrodes and pentodes makes possible a reduction in feedback and thus overcomes this difficulty. Sufficient oscillations are still generated because these tubes are more sensitive than triodes and require less grid voltage for satisfactory operation. The circuit connections for tetrode or pentode crystal-controlled oscillators are modified to provide the correct amount of feedback and to supply the necessary screen-grid voltage.

RESISTANCE-CAPACITANCE OSCILLATORS

Resistance-capacitance oscillators depend for their operation on the charge and discharge of a capacitor in series with a resistor, whereas inductance-capacitance oscillators depend on the resonant interchange of energy between a capacitor and an inductor. A better understanding of the action of resistance-capacitance oscillators may be gained through a brief review of the growth and decay of current in an R-C series circuit, as treated in texts on basic electricity.

Although there are various types of resistance-capacitance oscillators such as saw-tooth generators, multivibrators, blocking oscillators, and switching and counting circuits, only the basic saw-tooth generators and one of the typical multivibrator circuits are treated in this chapter. Other circuits will be treated as needed in advanced training courses.

Saw-Tooth Generator

Voltages having saw-tooth waveforms are widely used in television, radar, and many other electronic devices including test equipment. In each of these applications the saw-tooth wave is used to sweep the electron beam across the fluorescent screen of a cathode-ray tube.

NEON-TUBE SAW-TOOTH GENERATOR.—One of the simplest devices for developing this type of waveform is the gas-tube relaxation oscillator, shown in figure 7–17. Capacitor C is charged through resistor R until the potential across C reaches a value high enough to ionize the gas in the tube. Until this time the tube has a high impedance, but at the ionization potential its impedance drops to a low value and C discharges rapidly through it. When the voltage across C falls below the ionizing potential, the initial high impedance across the tube is reestablished and the capacitor stops discharging. Because the voltage across C is less than the value required to ionize the tube, the capacitor again charges.

For a given supply voltage, the frequency of the saw-tooth voltage depends upon the RC time constant and is varied by adjusting R.

A consideration of figure 7–17, B, indicates that the output voltage varies between the deionizing potential and the firing potential of the gas tube. The full B-supply voltage is not applied across C because the firing potential is a much lower value and the difference appears across R. Likewise, C does not completely discharge because when the deionizing potential is reached, C stops discharging. The capacitor voltage follows a normal RC charging curve between these two limits. The discharge follows a similar curve except that the discharge time is only a small fraction of the charge time because the resistance of the discharge path is only a small fraction of the resistance of the charge path and the curve is much steeper.

The output voltage curves for increased RC time constant and for increased B potential are shown in figures 7–18. For example, in figure 7–18, A, increasing the resistance, R,

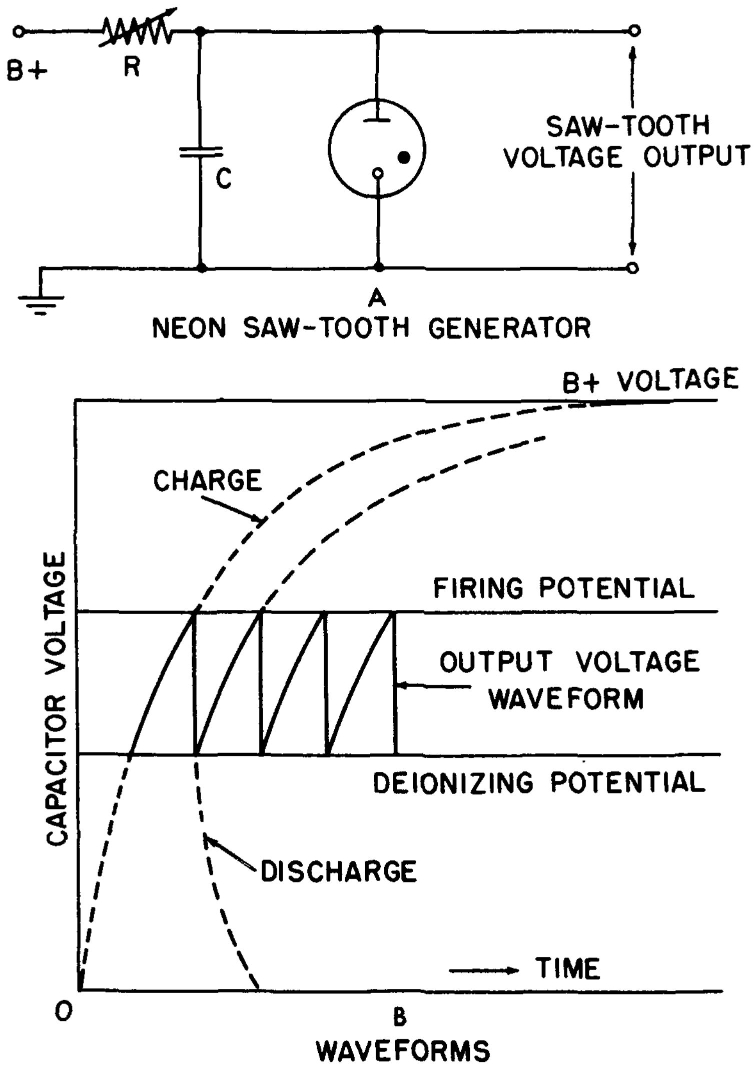
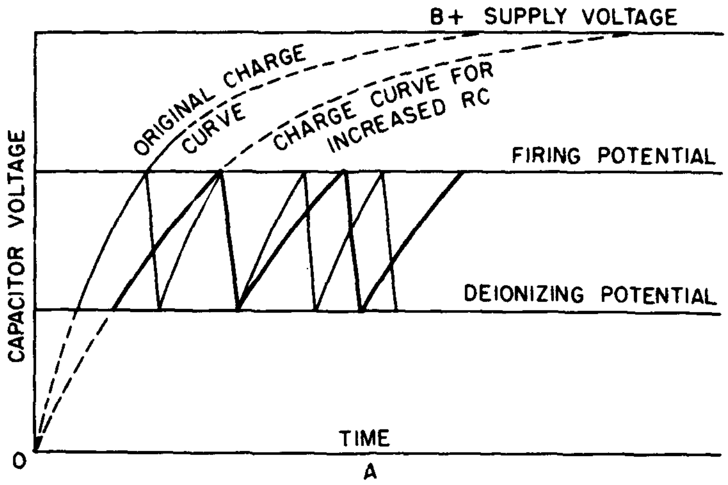
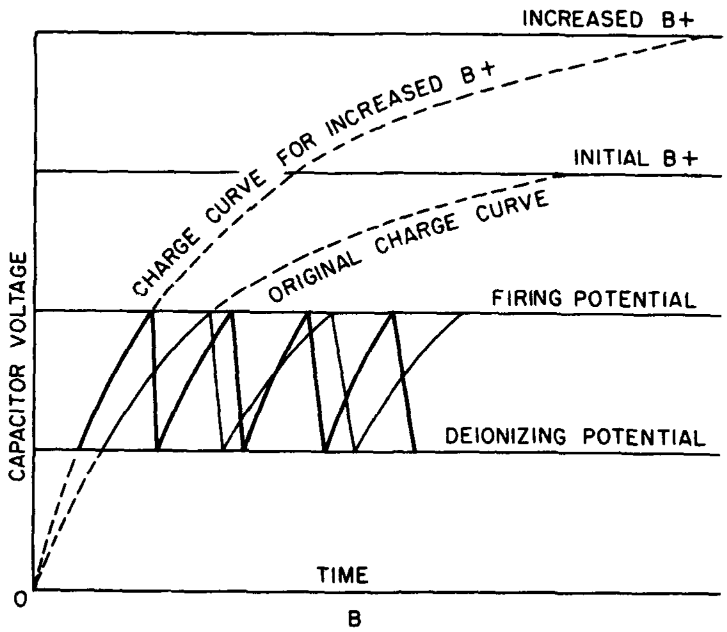


Figure 7-17.—Circuit of a neon saw-tooth generator.



A

R-C TIME CONSTANT INCREASED—FREQUENCY REDUCED



B

SUPPLY VOLTAGE INCREASED—FREQUENCY INCREASED

Figure 7–18.—Curves resulting from a variation in circuit constants.

increases the time necessary for C to charge to the ionizing potential, and the frequency is correspondingly decreased. Conversely, increasing the supply voltage (fig. 7-18, B) decreases the time necessary to charge C to the firing potential, and the frequency is correspondingly increased.

THYRATRON SAW-TOOTH GENERATOR.—The thyratron, or gas-filled triode, is generally used to produce saw-tooth waveforms and has certain advantages over the simple neon-tube saw-tooth generator. For example, the thyratron is more stable—that is, changes in the applied voltages or frequency do not alter its characteristics so readily. The deionizing time also is reduced.

The thyratron operates much the same as the neon tube except that the IONIZING potential is controlled by the grid. The DEIONIZING potential is affected very little by the grid bias. The more negative is the grid with respect to the cathode, the higher is the ionizing potential and the lower the frequency. A simple thyratron saw-tooth generator circuit and the output waveforms with high and low values of grid potential are shown in figure 7-19. The approximate frequency of the saw-tooth voltage is

$$f=\frac{1}{2.302\,RC\log_{10}\frac{E_b-E_2}{E_b-E_1}},$$

where f is the frequency in cycles per second, R the total charging resistance in megohms, C the total capacitance in microfarads, E_b the plate supply, E_2 the deionization potential, and E_1 the ionization potential in volts.

In this circuit the B-supply voltage of course must be larger than the ionizing potential of the tube. If a low potential is used, the output will have greater nonlinearity; in other words, a longer time will be required for capacitor C to become charged, and therefore the full charge curve is used. On the other hand, if a high voltage is used, only the lower, more linear, portion of the curve is utilized before the ionization potential is reached.

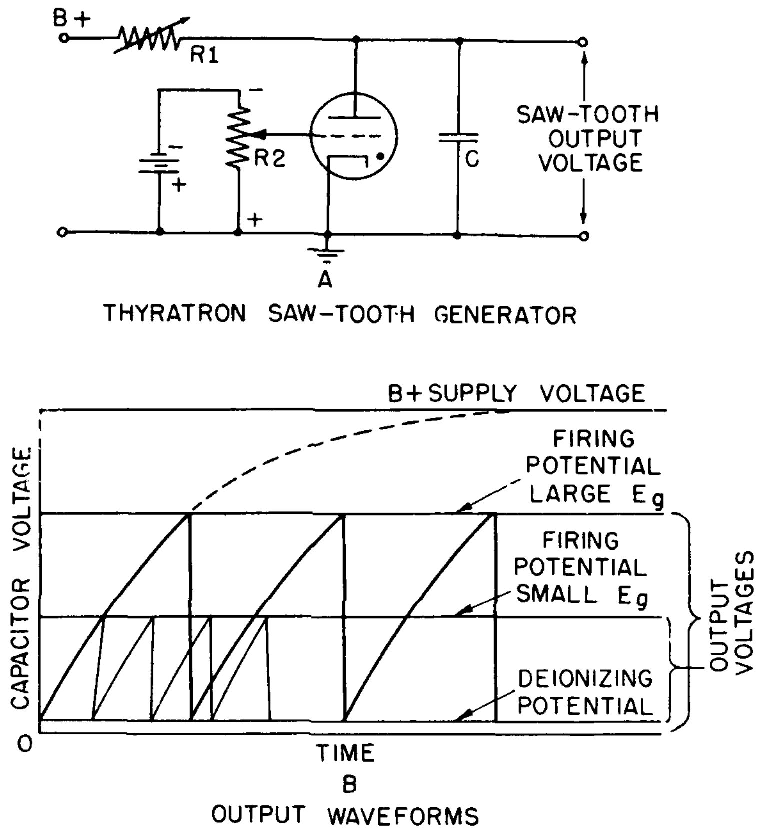


Figure 7–19.—Thyratron saw-tooth generator and output waveforms.

SYNCHRONIZED THYRATRON SAW-TOOTH GENERATOR.—A gas-tube relaxation oscillator does not produce oscillations that are stable in frequency. It can be synchronized with a constant frequency, however, by injecting a small voltage of the desired frequency into the grid circuit. A saw-tooth oscillator stabilized in this manner is shown in figure 7–20.

Circuit operation may be explained as follows: The oscillator is adjusted until its natural frequency is somewhat lower than that of the synchronizing signal, as indicated by

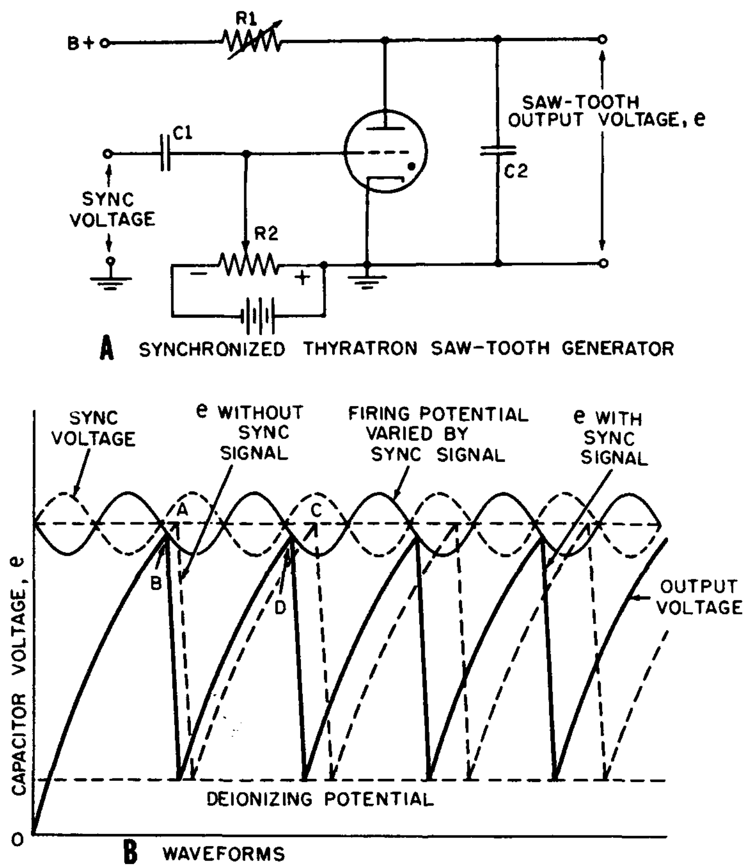


Figure 7–20.—Synchronized thyratron saw-tooth generator and waveforms.

the dotted saw-tooth curve (without sync signal) in figure 7–20, B. Without the sync signal the tube fires at points *A*, *C*, etc., but with the signal, the firing potential varies according to the instantaneous value of the grid potential. In other words, when the positive half of the sync signal is applied to the grid, the firing potential is reduced and the tube fires at points *B*, *D*, etc.; and, when the negative half

is applied, the firing potential is increased. If the synchronizing voltage is applied, the time for each oscillation is reduced from AC to BD, and the oscillator is locked to the frequency of the sync voltage. The oscillator may also be locked to a multiple or submultiple of the sync voltage.

The vertical and horizontal sweep oscillators (not necessarily thyratrons) in television receivers are typical circuits controlled by sync pulses sent out by the transmitter.

Multivibrators

A multivibrator is an electron-tube oscillator that utilizes two tubes or two sections of one tube to feed the output of one tube to the input of the other (and vice versa) by means of a resistance-capacitance coupling network. The output is essentially square wave, and for a FREE-RUNNING multivibrator, the frequency is determined by the values of R and C. The frequency may be easily controlled by the application of an externally generated signal to the circuit. When this type of signal is applied, the circuit is referred to as a DRIVEN multivibrator.

Before the multivibrator cycle of operation is analyzed, the following properties of electron-tube circuits are reviewed:

1. When the grid becomes less negative (more positive) with respect to the cathode, the plate current increases, and vice versa.
2. An increase in current through the plate load resistor causes a greater IR drop across it, and therefore the plate voltage is reduced. Conversely, lower plate current results in higher plate voltage.
3. There is a 180° phase shift between the grid signal voltage and the a-c component of plate voltage.
4. The voltage cannot build up or decay instantaneously across a capacitor.
5. Current flow through a resistor is from the negative end to the positive end.

6. A capacitor requires a definite time to charge or discharge through a resistor. The time necessary for a capacitor to charge to 63 percent or to discharge to 37 percent of its final voltage is known as the TIME CONSTANT of the circuit. Its value in seconds is equal to the product of the resistance in ohms and the capacitance in farads.

ECCLES-JORDAN TRIGGER CIRCUIT.—The basic multivibrator will be more easily understood if the action of the Eccles-Jordan trigger circuit is considered first. This type of circuit is used where an effect that is electrically the same as opening or closing a switch is desired. It is also used as a counting or scaling device.

In the strict sense of the word the Eccles-Jordan trigger circuit (fig. 7–21) is not an oscillator. It is instead a circuit

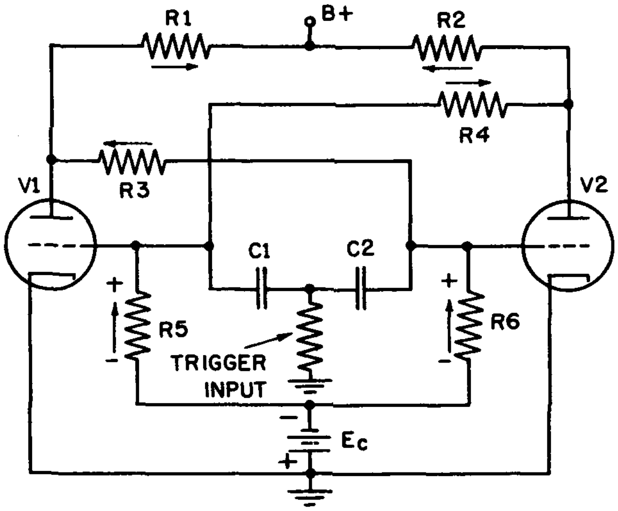


Figure 7–21.—Eccles-Jordan trigger circuit.

that has two conditions of equilibrium. One condition is when *V*1 is conducting and *V*2 is cut off, and the other is when *V*2 is conducting and *V*1 is cut off. Both conditions are quiescent—that is, there is no change in current or in any of the potentials until the tube is triggered by an ex-

ternal signal. When the trigger pulse is applied, the nonconducting tube conducts and the conducting tube ceases to conduct. On the next pulse, the reverse operation occurs. This circuit is thus aptly called a FLIP-FLOP CIRCUIT.

The grids of $V1$ and $V2$ are connected to voltage divider networks. The voltage divider network for $V1$ includes $R2$, $R4$, $R5$, and E_c. The network for $V2$ includes $R1$, $R3$, $R6$, and E_c. The voltage drops across $R5$ and $R6$ are individually less than E_c and because these voltages subtract from E_c, the grids are always negative with respect to the cathodes.

The action of the Eccles-Jordan circuit is as follows:

1. Assume that the cathodes are heated and that B voltage is applied to both tubes. If both tubes and their corresponding elements were exactly alike, equal currents would flow through the plate circuits. It is not likely, however, that the two tubes and their circuit elements would be balanced so exactly as to permit this to occur. Actually one tube starts to conduct an instant before the other or conducts more heavily than the other. Assume that $V1$ (fig. 7–21) conducts more current than $V2$.
2. The voltage drop across $R1$ is greater than the drop across $R2$, and the voltage at the plate of $V1$ is lower than the voltage at the plate of $V2$.
3. The lower voltage on the plate of $V1$ reduces the voltage across $R6$ and increases the negative bias on $V2$. The current of $V2$ is further reduced.
4. Therefore the voltage at the plate of $V2$ is increased, and in turn the drop across $R5$ is increased, reducing the negative bias on the grid of $V1$. Current in the plate circuit of $V1$ is further increased and the plate voltage is further decreased.
5. The action is cumulative, and very quickly a condition is reached when the plate current of $V1$ reaches a maximum and the plate current of $V2$ is cut off. This is one condition of stable equilibrium. During this quiescent period the drop across $R5$ is larger than the drop across $R6$.

6. Assume that a positive-going signal is applied simultaneously to the grids of both tubes at the trigger input. Since *V*1 is already passing a heavy current, the positive pulse on its grid has little effect on the flow of current through the tube. Tube *V*2, however, is cut off and the positive pulse on its grid, if of sufficient amplitude, removes the negative bias momentarily. Current then flows in the plate circuit of *V*2.
7. Plate voltage of *V*2 is reduced, and the reduced voltage across *R*5 makes the grid of *V*1 more negative.
8. Plate current in *V*1 is reduced and the plate voltage of *V*1 is increased.
9. The increased voltage of *V*1 increases the drop across *R*6, further reducing the bias on *V*2, and its plate current continues to increase, thus applying more negative bias to *V*1.
10. Current in *V*1 quickly ceases, and at the same time current in *V*2 reaches saturation.

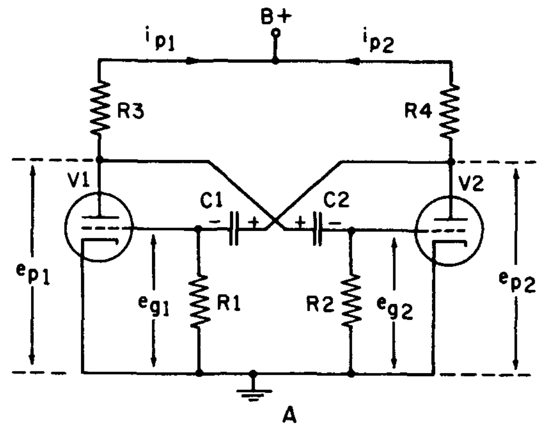
If negative-going pulses had been used, the conducting tube would have been the first one affected. Its plate current would have been decreased, with the same end result (*V*1 cut off and *V*2 conducting). One alternation is thus completed for each pulse, and two pulses are necessary to complete a full cycle.

Basic free-running multivibrator.—The basic free-running multivibrator circuit is shown in figure 7–22. Such a vibrator circuit is simply a two-stage *R-C* coupled amplifier with the output of the second stage coupled through *C*1 to the input of the first stage and the output of the first stage coupled through *C*2 to the input of the second stage.

Because the voltage that is fed back in each case is of the proper polarity to reinforce the voltage on the grid of the tube receiving the feedback voltage, signals are reinforced and oscillation takes place.

The operation of the basic free-running multivibrator shown in figure 7–22, A, may be summarized as follows:

When the cathodes are heated and the plate potential is



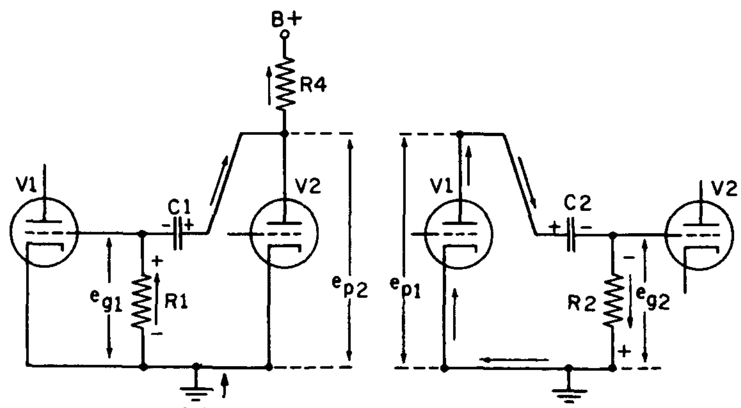
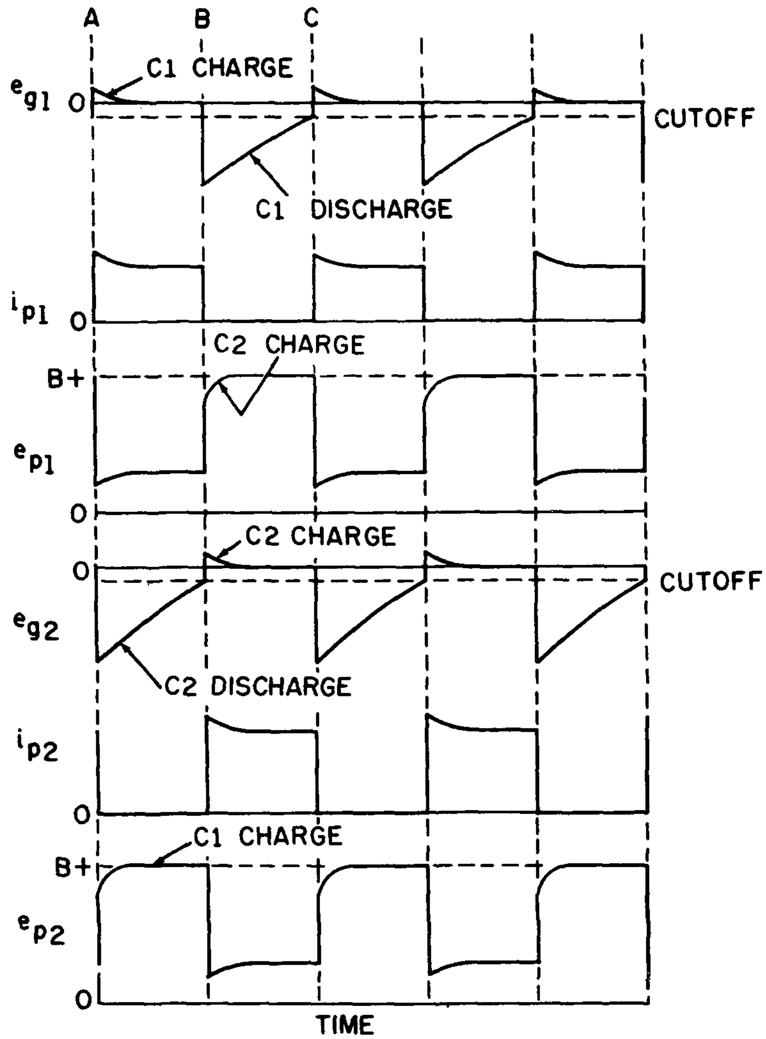


Figure 7-22.—Basic free-running multivibrator and waveforms.

applied, both tubes begin to conduct. Initially the plate currents are nearly equal, but there is always a difference between them. The slight initial unbalance brings about a cumulative or regenerative switching action, which in this example is assumed to end with i_{p1} increased to a maximum



C
WAVEFORMS

Figure 7-22.—Basic free-running multivibrator and waveforms—Continued

value of i_{p2} reduced to zero. Although described as if it occurred slowly, this switching occurs with extreme rapidity—in a fraction of a microsecond in a well-designed multivibrator. This action is followed by a relatively long period in which the tubes are quiescent. During this interval one capacitor charges and the other discharges.

Assume that initially i_{p1} rises more rapidly than i_{p2}. Plate voltage e_{p1} falls (because of the increased drop in $R3$) and $C2$ discharges through $R2$, making the grid of $V2$ negative, thus reducing i_{p2}. Plate voltage e_{p2} rises (because of the decreased drop across $R4$) and $C1$ charges through $R1$, thus applying a positive bias to the grid of $V1$. The plate current of $V1$ rises to a maximum value, and $V2$ is cut off.

The charge path for $C1$ and the discharge path for $C2$ are shown in figure 7–22, B. The waveforms of plate current and plate and grid voltages are shown in figure 7–22, C.

The negative grid voltage applied to $V2$ results from the discharge of $C2$ through $R2$ and returns to zero as the capacitor discharge is completed. When the bias is reduced to the cutoff-point plate current i_{p2} begins to flow, and a second switching action takes place. This switching action is like the first except that i_{p2} is increasing and i_{p1} is decreasing. Plate voltage e_{p2} decreases (because of the increased drop across $R4$) and $C1$ discharges through $R1$, making the grid of $V1$ negative, thus reducing plate current i_{p1}. Plate voltage e_{p1} rises (because of the decreased drop across $R3$) and $C2$ charges through $R2$, making the grid of $V2$ positive. Thus the second switching action ends with $V2$ carrying maximum current and $V1$ cut off.

During the cycle of operation, current is maintained at a relatively steady value in one tube during the interval that the other tube is cut off. The action repeats continuously with first one tube and then the other conducting.

The frequency of the oscillations depends on the time constants of the coupling net works, R_1C_1 and R_2C_2. For a uniform waveform, $R_1=R_2$ and $C_1=C_2$. The frequency is expressed as

$$f=\frac{1,000}{R_1C_1+R_2C_2}=\frac{1,000}{2R_1C_1},$$

where f is in kilocycles, R_1 in ohms, and C_1 in microfarads.

MULTIVIBRATOR SYNCHRONIZED BY SINE WAVES.—Because free-running multivibrators have poor stability they are often synchronized with another frequency that forces the period of the multivibrator oscillation to be exactly the same as that of the synchronizing frequency. Such a multivibrator is said to be driven by the synchronizing voltage.

Sine waves or pulses are generally used for synchronizing purposes, although waveforms of almost any shape could be used. Synchronization by means of sine waves will be considered first. The synchronizing signal may be injected at the cathode or between the grid and cathode; however, only cathode injection is considered in this chapter.

Figure 7–23 shows a multivibrator with the synchronizing sine-wave voltage, e_k, applied to the cathode. A set of voltage curves is also shown in order to clarify the discussion.

The actual grid-to-cathode voltage of $V1$, which is the voltage controlling the flow of plate current, is the algebraic sum of e_{g1} and e_k. The source of the sinusoidal synchronizing voltage should have a low internal impedance so that the flow of i_{p1} through this source will not bring about an alteration in the shape of the wave.

The operation of the circuit may be explained as follows:

If the multivibrator is properly balanced and running freely, the voltage curve of e_{g1} is that shown between time A and time B in figure 7–23. Because there is no synchronizing voltage on the cathode, its voltage remains constant at ground potential.

At time B the synchronizing voltage is applied and e_k begins to vary sinusoidally. Although e_{g1} is not affected by this variation, the grid-to-cathode potential now contains this sinusoidal voltage component. Thus, the effective cutoff voltage of the tube varies sinusoidally about its normal cutoff value in phase with the synchronizing voltage on the cathode. The cathode voltage curve, e_k, and the effective

cutoff voltage curve, e_{co}, are shown with the e_{g1} curve in order to explain the synchronizing action.

The instant at which $V1$ conducts occurs when the e_{g1} curve crosses the e_{co} curve. At instant B, e_k starts to rise in a positive direction and i_{p1} is decreased. The positive-going voltage produced at the plate of $V1$ initiates the switching action via $C2$ to the grid of $V2$ and from the plate of $V2$ via $C1$ back to the grid of $V1$. Then e_{g1} drops along line BC instead of along DE, as it would in the free-running state, and $V1$ is quicklv cut off. $C1$ discharges along curve CFG;

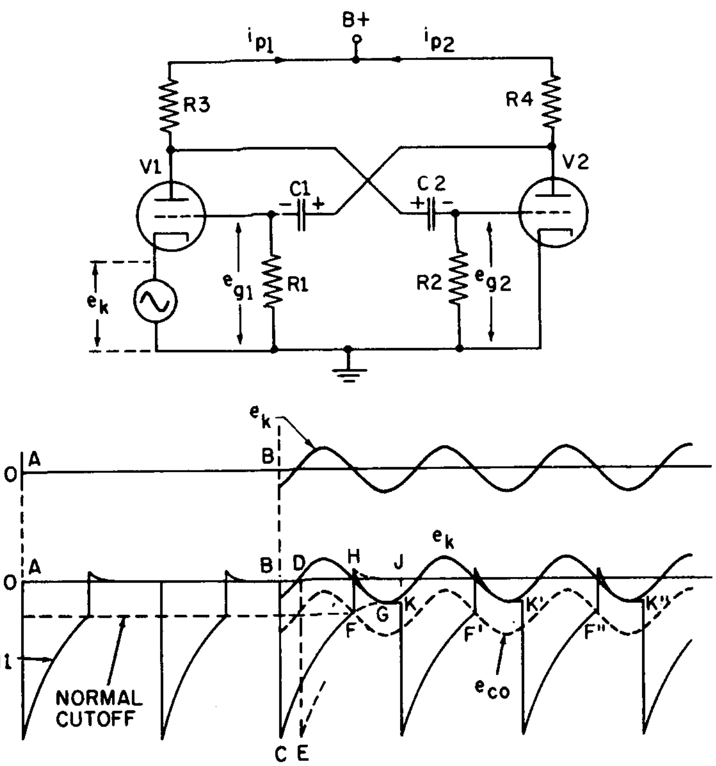


Figure 7–23.—Multivibrator with sine-wave synchronizing voltage applied to the cathode.

but since the e_{g1} curve intersects the e_{c0} curve at F, the switching action by which $V1$ is made conducting and $V2$ is cut off takes place at F, instead of at G as it would in the free-running state.

The switching action drives the grid of $V1$ positive, but the resulting grid current quickly charges $C1$, and the grid returns to cathode potential. The grid voltage, e_{g1}, does not follow curve HJ, as it would in the free-running state. Instead, the grid draws current because the synchronizing voltage causes the cathode to be negative with respect to ground at this time, and e_{g1} follows the cathode voltage along curve HK. When the cathode voltage begins to rise in a positive direction the plate current of $V1$ starts to decrease. At instant K the rise in voltage at the plate of $V1$, resulting from the decrease in i_{p1}, is large enough to drive $V2$ into conduction, and the tubes are rapidly switched.

The action of the sine-wave synchronizing voltage forces the time of one cycle to be shorter and the frequency to be higher than it would be without the synchronizing signal. Switching in one direction occurs at instants F, F', F'', etc., and switching in the other direction occurs at instants K, K', K'', etc. With the exception of the short transition time, the period of the multivibrator is equal to the period of the synchronizing voltage. Thus, the multivibrator is said to be synchronized.

The synchronizing voltage can make the multivibrator operate above or below its natural frequency. However, if an attempt is made to pull the multivibrator to a frequency that is too high, it will synchronize at a frequency that is one-half or some other division of the synchronizing frequency, and FREQUENCY DIVISION may be obtained.

MULTIVIBRATOR SYNCHRONIZED BY PULSES.—Multivibrators may be synchronized also by short positive or negative trigger pulses. Figure 7–24 shows the effects of positive pulses on the multivibrator grid-voltage waveform.

A positive pulse insufficient to drive the grid above cutoff applied to a nonconducting tube at instant A does not cause switching action. The only effect is to reduce the

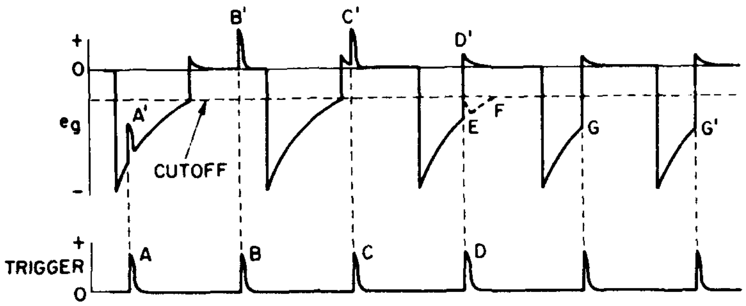


Figure 7-24.—Waveforms on the synchronized grid of a multivibrator driven by positive pulses.

negative bias slightly, as shown at A′. A positive pulse applied to a tube that is already conducting (points B or C) serves only to increase momentarily the grid voltage and thus to increase the plate current momentarily. It has no effect on the action of the multivibrator.

With the exception of the variations in the e_g waveform at times A', B', and C' the multivibrator is essentially free running. If applied at instant D, however, the positive trigger pulse is sufficient to overcome the negative voltage on the grid and drives the grid above cutoff. The cycle of the multivibrator is thereby shortened by an amount EF.

For proper synchronization, the natural period of the free-running multivibrator must be greater than the time interval between pulses. Under these circumstances the positive trigger pulses cause the switching action to occur earlier in the cycle than it would in the free-running state. Thus, the tube conducts at E, G, and G', whereas it would have conducted later in each instance had the pulses not been applied. Under these circumstances the frequency of the multivibrator is forced to become the same as the repetition frequency of the trigger pulses.

The multivibrator may be synchronized to a submultiple of the trigger frequency if both frequencies are such that every second, third, fourth, etc., synchronizing pulse occurs at the right time so that it will drive the grid voltage of the nonconducting tube above cutoff.

The multivibrator may be used as a sweep-frequency generator for cathode-ray tube applications—as in television, where magnetic deflection is commonly used. The horizontal and vertical oscillators in the receiver are triggered by pulses sent out from the transmitter so that they will be locked in step with similar oscillators at the transmitter. The multivibrator may be used as a source of square waves, as an electronic switch for various applications, or as a means of obtaining frequency division. It is often used to introduce a time delay between the operation of two circuits by using the leading edge of the square wave to trigger one circuit and the trailing edge to trigger another. The time delay can be controlled by varying the *RC* time constants of the multivibrator circuit.

In television transmitters and in radar the action of the multivibrators is accurately timed by triggering them with pulses from a master oscillator circuit.

When used as an electronic switch, one multivibrator tube allows its associated amplifier to function normally while the second multivibrator tube holds its amplifier cut off, and vice versa. In radar, multivibrators are used principally as electronic switches to produce GATE VOLTAGES that permit a part of a circuit to operate only during an accurately controlled time interval.

QUIZ

1. What are two conditions necessary to produce sustained oscillations in an electron-tube oscillator?
2. What determines the upper frequency limit of electron-tube oscillators?
3. In a tickler-feedback oscillator what grid circuit test may be made to indicate proper operation?
4. How is energy coupled from the plate circuit back into the tuned grid circuit in the series-fed Hartley oscillator shown in figure 7–3?
5. In a series-fed Hartley oscillator what action maintains the oscillations in the tank circuit when the plate current is zero and no energy is being supplied to the oscillator circuit?
6. What is the advantage of self-bias in a series-fed Hartley oscillator?
7. What circuit arrangement distinguishes the shunt-fed Hartley oscillator from the series-fed type?

8. How does the voltage divider in the Colpitts oscillator tank circuit differ from the one used in the Hartley oscillator?
9. How is feedback obtained in a tuned-plate tuned-grid oscillator?
10. Which tuned circuit in terms of the relative tank circuit Q (high or low) in a TPTG oscillator determines the oscillator frequency?
11. Push-pull oscillators are generally used in what frequency ranges?
12. The electron-coupled oscillator combines the function of oscillator and amplifier in which respective circuits (fig. 7–9)?
13. What effect does increasing the plate voltage have on the frequency of an electron-coupled oscillator?
14. How is the effect of negative resistance produced in the transitron oscillator of figure 7–10?
15. What are the advantages of quartz over Rochelle salt when used as oscillator crystals?
16. What is the difference between the X-cut and the Y-cut crystal as far as the temperature effect on frequency is concerned?
17. What type of cut has a nearly zero temperature coefficient?
18. How is a high degree of frequency stability maintained in the oscillation section of a standard broadcast transmitter with respect to temperature changes?
19. Why do oscillations cease in the crystal-controlled oscillator of figure 7–15 when the capacitance is increased beyond the point where the dip in the plate-current tuning curve occurs?
20. Why are tetrodes or pentodes frequently used in crystal oscillators?
21. For a given supply voltage upon what does the frequency of a neon-tube saw-tooth generator depend?
22. What effect does increasing the B-supply voltage have on the frequency of a neon-tube saw-tooth generator?
23. Name two advantages of the thyratron saw-tooth generator over the neon-tube saw-tooth generator.
24. What is the advantage of using a high B-supply voltage on a thyratron saw-tooth oscillator?
25. Sketch an Eccles-Jordan trigger circuit.
26. What determines the frequency of a free-running multivibrator?
27. Why are multivibrators often synchronized with another frequency?
28. At what two positions in the multivibrator circuit are sync signals commonly injected?
29. What is the result of "pulling" the multivibrator frequency to a value that is too high?
30. What is the principal use of multivibrators in radar sets?

CHAPTER 8

MODULATION AND DEMODULATION

INTRODUCTION

MODULATION is the process by which the amplitude or frequency of a sine-wave voltage (the CARRIER) is made to vary with time according to the voltage or current variations of another (MODULATING) signal. The carrier is usually of a higher frequency than the modulating signal.

DEMODULATION, OR DETECTION, is the process by which the audio signal is recovered from the carrier at the receiver.

Audio frequencies extend from 15 cycles per second to 20,000 cycles per second, as shown by the audio-frequency spectrum in figure 8–1. The human voice extends from about 87 cps to 1,175 cps. The violin has a range of from about 200 cps to 3,000 cps, and the bass viol extends from about 40 cps to 250 cps. The pure tones of the piccolo extend to about 5,000 cps. However, combinations of sound frequencies produce harmonics that extend up to 20,000 cps. These combinations of frequencies give to the speech or music the identifying characteristics that distinguish one person from another and one type of musical instrument from another.

For the following reasons it is not practical to transmit electromagnetic waves at audio frequencies—that is, without the use of a modulated r-f carrier: (1) The range of such transmission would be very limited because of the poor radiation efficiency of antennas at these low audio frequen-

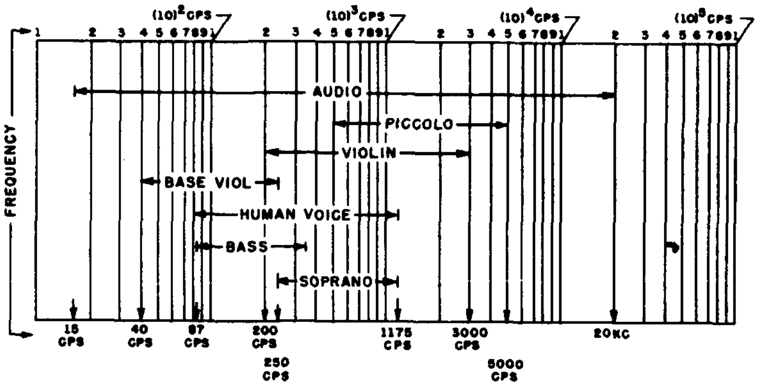


Figure 8–1.—Audio-frequency spectrum.

cies; (2) all transmitters would operate in the same frequency range, and therefore the signals could not be separated in the receivers; (3) the antennas would have to be excessively long to be in resonance at the middle frequency in the audio range (it follows that the antenna would be considerably out of tune at the end frequencies); and (4) the inductors and capacitors would have to be very large in order to produce resonance at these low frequencies.

These problems may be overcome by the use of a modulated r-f carrier at the transmitter end and a method of removing the modulation at the receiver end. The efficiency of radiation is thereby improved; each carrier with its associated modulation component is confined to a relatively narrow band in the r-f spectrum; and interference (while still a problem in certain instances) is not intolerable.

AMPLITUDE MODULATION

Amplitude modulation (a-m) may be defined as the variation of the strength of the r-f output of a transmitter at an audio rate. In other words, the r-f energy is made to increase and decrease in power according to the audio (sound) frequencies. If the audio frequency is high, the radio

frequency must vary in amplitude more rapidly than if the audio frequency were low. If the a-f tone is loud in volume, the r-f energy must increase and decrease by a larger percentage than if the a-f tone were soft. Thus, the r-f variation must correspond in every respect with the a-f variations.

Figure 8-2 indicates that the resultant wave with single-

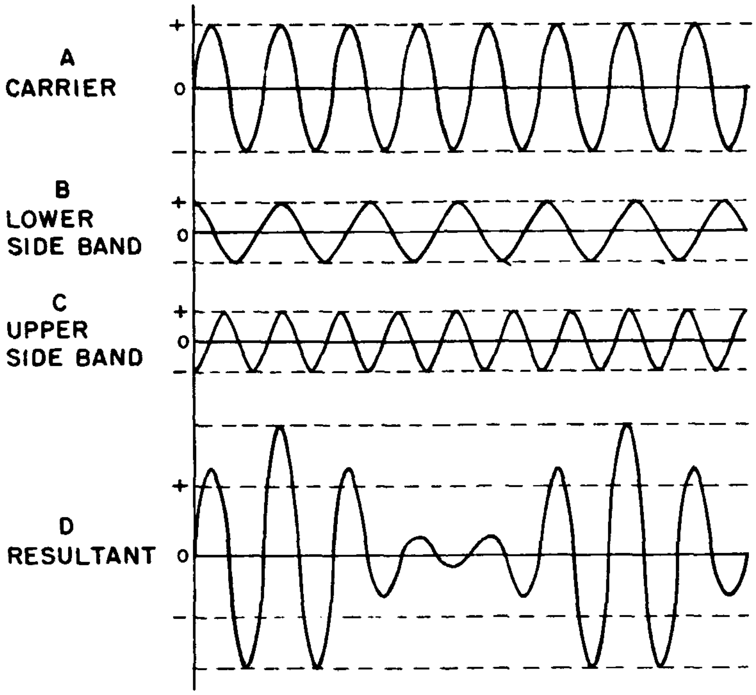


Figure 8-2.—Curves showing how the carrier and side bands combine to produce an a-m wave with 100-percent modulation.

tone amplitude modulation consists of three separate waves. The carrier with no modulation is shown in figure 8-2, A. The lower side band has a frequency equal to the difference between the modulation and carrier frequencies and is shown in figure 8-2, B. The upper side band has a frequency equal to the sum of the carrier and modulation frequencies and is

shown in figure 8–2, C. The carrier and the side bands are not merely a mathematical abstraction; they may be separated from one another by filters and used individually.

In an a-m wave only the side bands contain the intelligence to be transmitted; the audio frequency as such is not transmitted. Because the modulating frequencies are the information to be transmitted, as much power as possible should be put into the side bands. In other words, the amplitude of the modulated carrier wave should be varied as much as possible. When the amplitude is carried completely to zero during the modulation cycle, as in figure 8–2, D, the modulation is 100 percent; and the side bands contain the maximum permissible amount of power, or one-half the carrier power, because modulation greater than 100 percent causes distortion.

Bandwidth Requirements

Because an a-m wave has side bands on each side of the carrier, the transmission of information by amplitude modulation requires the use of a band of frequencies rather than a single frequency. Music may contain frequency components as high as 15,000 cps, so that music modulated upon a carrier produces side-band components extending to 15,000 cycles on each side of the carrier frequency.

Local a-m broadcast stations are allocated a total bandwidth of only 10 kc (5 kc on each side of the carrier frequency) because of the large number of stations on the air. Since the total bandwidth is only 10 kc, audio frequencies above 5 kc cannot be transmitted without causing interference between stations.

Naval a-m communication sets operate within the relatively narrow bandwidth required by voice or code transmission. The RBB and RBC receivers utilize a narrow, adjustable bandwidth of only a few tenths of 1 percent of the carrier frequency, as the selectivity curves of the RBC receiver (fig. 8–3) show. A narrow bandwidth permits the use of more selective circuits. These selective circuits reduce noise and the number of interfering signals.

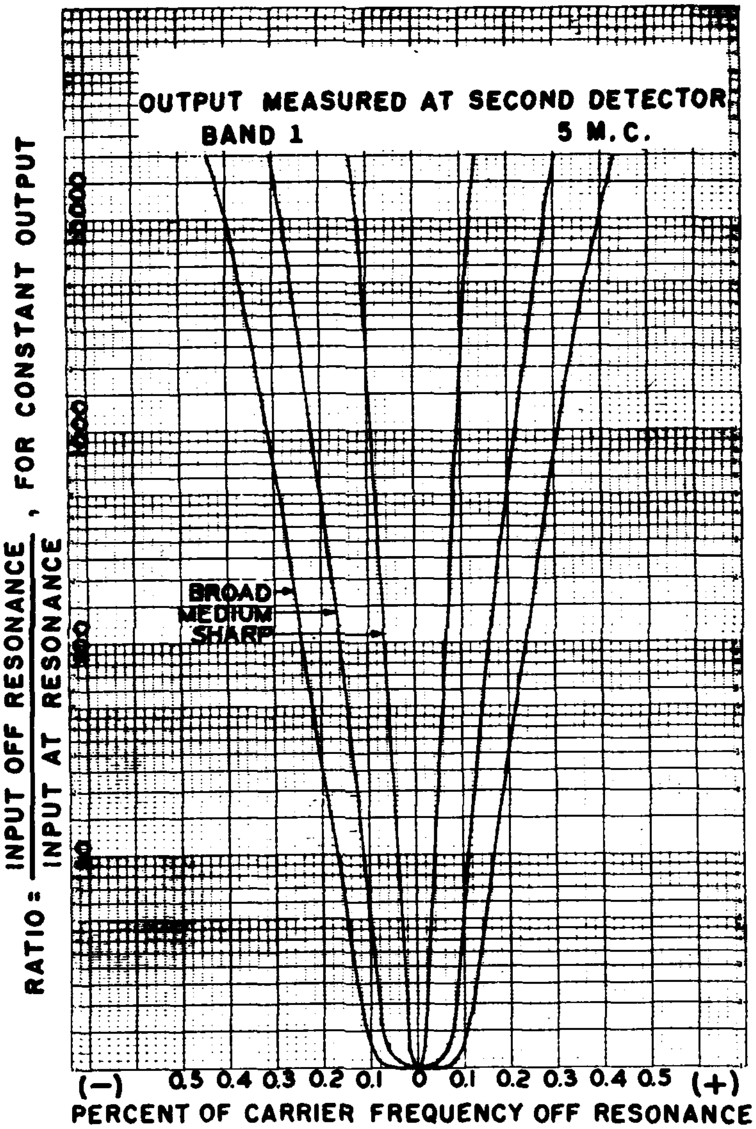


Figure 8–3.—Selectivity curves of the RBC radio receiver.

Percentage of Modulation

The degree of modulation in an a-m wave is expressed by the percentage of maximum deviation from the normal value of the r-f carrier. The effect of such a modulated wave, as measured by receiver response, is proportional to the degree, or percentage, or modulation.

An a-m wave is shown in figure 8-4. The percentage of modulation may be determined by the equation

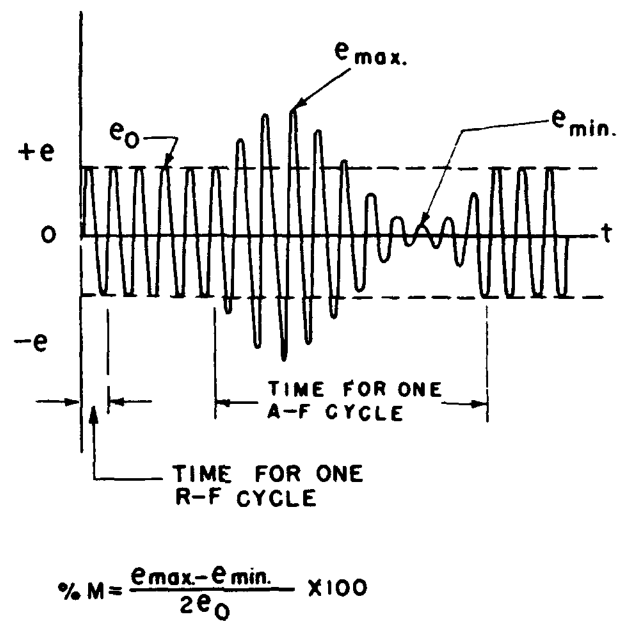


Figure 8-4.—Percentage of modulation.

$$\text{percentage of modulation}=\frac{e_{max}-e_{min}}{2e_o}\times 100,$$

where e_{max} is the maximum instantaneous value of the r-f voltage across the transmitter tank circuit, e_{min} the minimum instantaneous value of the r-f voltage, and e_o the maximum instantaneous value of the r-f voltage in the absence of modulation.

It is important that the amplitude be varied as much as possible, because the output of a detector in a radio receiver varies with the amplitude variations of the received signal. Thus a comparatively low-powered, but well modulated, transmitter often produces a stronger signal at a given point than a much higher powered, but poorly modulated, transmitter located the same distance from the receiver.

If modulation exceeds 100 percent there is an interval during the audio cycle when the transmitter is removed completely from the air. For example, in figure 8–5, the

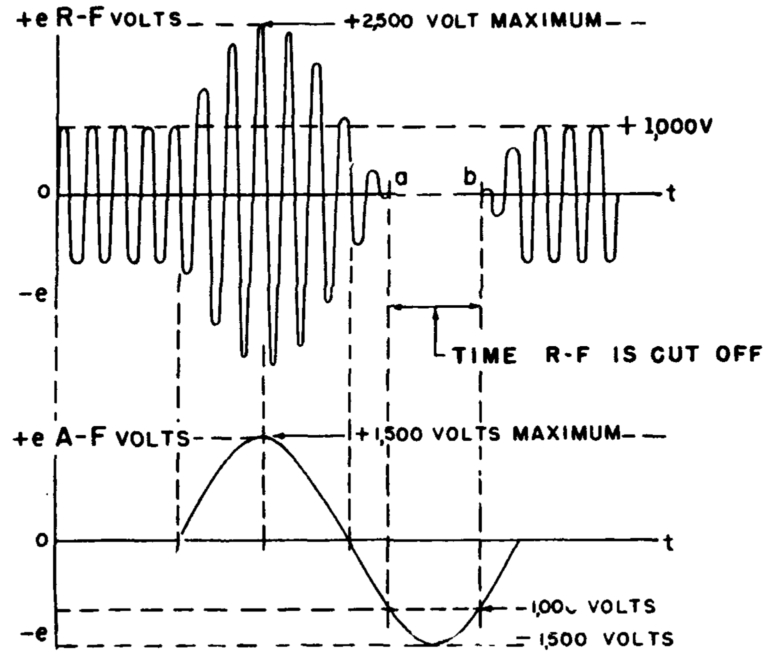


Figure 8–5.—Amplitude modulation in excess of 100 percent.

modulation is shown in excess of 100 percent. The audio voltage is assumed to have a peak value of 1,500 volts, and the unmodulated carrier voltage has a peak value of 1,000 volts. A more detailed analysis of overmodulation is included later in this chapter under "Plate Modulation."

Systems of Amplitude Modulation

A block diagram of an a-m radiotelephone transmitter is shown in figure 8–6. The top row of blocks indicates the r-f

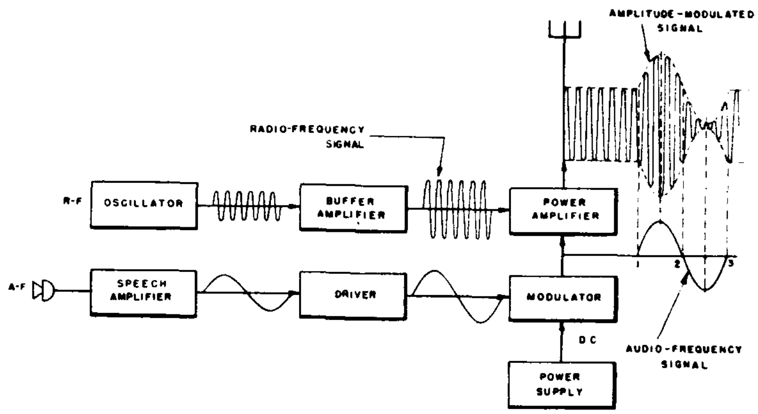


Figure 8–6.—A-m radiotelephone transmitter.

section. The next row of blocks indicate the a-f section; and the lower block indicates the power supply, which provides all d-c potentials.

The r-f section generates the high-frequency carrier radiated by the antenna. The methods by which the r-f signal is generated are treated in chapter 7.

The a-f section includes a speech amplifier that receives a few millivolts of a-f signal from the microphone (chapter 9) and builds it up to several volts at the input to the driver stage. This stage is made up of power amplifiers (chapter 6) that convert the signal into a relatively large voltage and appreciable current at the input to the modulator. The modulation transformer is capable of handling considerable audio power. Its output is fed to the final r-f power amplifier in such a way as to alternately add to and subtract from the plate voltage of the r-f amplifier.

The result is that the amplitude of the r-f field at the antenna is gradually increased in strength during the time

the a-f output is increasing the r-f power and gradually decreased in strength during the time the a-f output is decreasing the r-f power.

In other words, during the positive alternation of the audio signal (between point 1 and point 2 in fig. 8–6), the amplitude of the r-f output wave is increased, and during the negative alternation (between point 2 and point 3) it is decreased. Amplitude modulation consists of varying the amplitude of the r-f antenna current (and r-f output wave) gradually over the relatively long a-f cycle. Thus, the r-f field strength is alternately increased and decreased in accordance with the a-f signal and at the a-f rate.

There are a number of methods of producing amplitude modulation, such as plate modulation, grid modulation, screen-grid modulation, and so forth, but the two most important are plate modulation and grid modulation.

PLATE MODULATION.—One method of accomplishing amplitude modulation, called HIGH-LEVEL PLATE MODULATION, is shown in figure 8–7. It is called high-level plate modulation

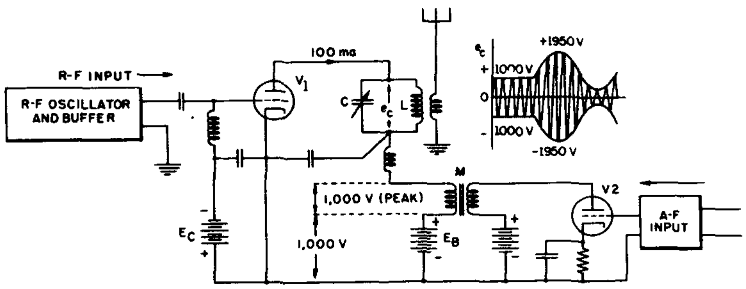


Figure 8–7.—High-level plate modulation.

because the audio signal is injected in the plate circuit at a high level of plate voltage. The triode class-C power amplifier, *V*1, is tuned to a resonant carrier frequency of 1 megacycle. The plate current is 100 milliamperes, and the plate voltage supply is 1,000 volts. The grid input driving voltage is produced by the r-f oscillator and buffer amplifier (used

to isolate the oscillator from the power stage), shown by block diagram at the left of figure 8–7.

An audio signal of approximately 1,000 volts (peak) having a sine waveform and a frequency of 1,000 cps is produced by the a-f section shown by block diagram at the right of figure 8–7 and appears as an output of the modulation transformer, *M*. This audio output voltage is in series with the r-f voltage across the tank circuit, *LC*, and the plate power supply, B+. In this example, conditions are established for the modulation of the 1-mc r-f output voltage with a single a-f signal having a sine waveform.

The 1-mc r-f carrier wave combines with the a-f signal to form side bands. The frequency of the r-f voltage developed across the plate tank is 1,000 times the audio frequency. Thus, the time for one complete audio cycle is long enough to include 1,000 cycles of r-f energy in the tank circuit, *LC*.

Before an audio-modulation signal is introduced, the r-f signal applied to the grid of *V*1 causes the triode to conduct periodically. During each conducting period, capacitor *C* charges. When the grid swings below cutoff, the triode stops conducting and the capacitor discharges through the coil.

The exchange of energy between the coil and capacitor accounts for the a-c voltage developed across the tank. In this exemple, the peak a-c voltage across the tank capacitor, *C*, which received its charge from the 1,000-volt B supply, is approximately 1,000 volts, neglecting losses.

Plate voltage varies between 1,000 ± 1,000, or 2,000 volts and 0 volts, as the capacitor voltage varies between +1,000 and −1,000 volts. Plate voltage is above the B-supply voltage during that part of the cycle when the grid voltage is below cutoff and the triode is not conducting. Plate voltage is below the B-supply voltage during the time the grid is above cutoff and the triode is conducting. Energy to supply the tank circuit losses comes from the B supply, and the flywheel effect in the tank circuit accounts for the sine waveform of r-f voltage and current within the tank.

Now assume that the a-f voltage of sine waveform is

introduced at M in series with the r-f tank and the B supply. Consider that portion of the a-f cycle in which the voltage is gradually rising according to the sine-wave variation so that the polarity aids the voltage of the B supply. For 250 cycles of radio frequency (one-quarter of one a-f cycle), the total voltage available for charging capacitor C in the plate-tank circuit is increasing from 1,000 volts to a maximum of 2,000 volts. Assume that the capacitor charges to 2,000 volts. The plate voltage at the end of 250 r-f cycles is a maximum of 2,000+2,000, or 4,000 volts. Thus, with 100-percent modulation, and on the positive audio peaks, the value of the peak-to-peak voltage across capacitor C is approximately double the peak-to-peak value it would have without modulation, or 4 times the B-supply voltage. The breakdown voltage across the capacitor restricts the r-f power output permissible with voice modulation.

The tank voltage starts to decrease between the 250th r-f cycle and the 500th r-f cycle as the a-f voltage falls from a maximum of 1,000 volts to zero volts during this period. From the 500th to the 750th r-f cycle the polarity of the a-f voltage is reversed, and reaches a maximum negative value at the 750th r-f cycle, and the tank capacitor charges up to the difference voltage between the 1,000-volt B supply and the instantaneous a-f value. At the instant the a-f voltage is maximum negative (−1,000 volts, at the 750th r-f cycle), the tank capacitor is not charging at all. The tank voltage is zero, and the r-f output energy momentarily becomes zero. Then the a-f voltage starts to fall from maximum negative to zero and the opposition that it offers to the power-supply voltage (B+) is reduced. The plate-tank capacitor starts charging again, and the r-f tank current again gradually increases during the subsequent 500 r-f cycles of operation.

The transmitting antenna is coupled to the tank, and the instantaneous antenna current increases and decreases in accordance with the tank-current variations. An r-f ammeter in the antenna circuit indicates the effective value of the current and not instantaneous values.

To understand the effective current indication corresponding to an a-m signal, study the distribution of r-f and a-f power in the tank circuit of figure 8–7. Before any a-f signal is injected, the tank circuit is assumed to be resonant. Plate current is 100 milliamperes and is composed of a rapid succession of nonsinusoidal pulses due to class-C operation. A d-c milliammeter in the plate circuit indicates a value that approaches the maximum value of these pulses. The momentum of its moving coil is too great to allow it to fall below the maximum indication in the brief interval between the r-f pulses. The power supplied to the r-f tank comes chiefly from the plate power-supply source. This source supplies 1,000 volts d-c and a maximum of 100 milliamperes, or a power supply of 100 watts. The peak value of the a-f voltage introduced at M is assumed to be 1,000 volts, which produces in this instance 100-percent modulation. The a-f voltage is assumed to be of a sine waveform; and the peak value of the current is assumed to be 100 milliamperes. The current is the same in all parts of a series circuit, and the output winding of the modulation transformer, M, is connected in series between the tank and B+. The a-f power supplied to this circuit is equal to one-half the product of the maximum voltage and the maximum current. Thus, the power supplied by the a-f modulator is

$$\frac{1{,}000 \times 0.1}{2} = 50 \text{ watts.}$$

The transmitter output power (neglecting losses) before modulation is 100 watts, and after modulation is 150 watts. If the equivalent antenna load resistance is assumed to be 100 ohms, the antenna current before modulation is

$$I = \sqrt{\frac{P}{R}} = \sqrt{\frac{100}{100}} = 1 \text{ ampere.}$$

When modulation occurs, the power in the preceding example increases to 150 watts. The antenna current is then

$$I=\sqrt{\frac{P}{R}}=\sqrt{\frac{150}{100}}=1.224 \text{ amperes},$$

which is an increase of 22.4 percent.

In the preceding example the modulation is 100 percent, which is the condition that exists when the audio-power input is equal to one-half the r-f power input. When 100-percent modulation occurs, antenna current increases 22.4 percent above the unmodulated value.

As previously stated, if the modulation exceeds 100 percent, there is an interval during the audio cycle when the transmitter output is zero. For example, in figure 8–5 the modulation is shown in excess of 100 percent. Assume that the audio input voltage of figure 8–7 has a maximum value of 1,500 volts instead of 1,000 volts. At the instant the audio voltage is maximum and adds to the power-supply voltage, the r-f voltage across the plate-tank circuit rises to 1,000+1,500, or 2,500 volts.

On the negative alternation of the audio cycle, when the audio voltage subtracts from the power-supply voltage, the plate voltage reverses and becomes negative with respect to ground during the interval *a* to *b* (fig. 8–5) when the audio voltage exceeds the voltage of the power supply. Tank-circuit oscillations cease during this interval because power cannot be supplied to the tank circuit while the plate of the triode is negative. This condition is called OVERMODULATION and results whenever the audio-modulation voltage exceeds the d-c voltage of the power-supply circuit. Overmodulation not only produces a distorted envelope, but also produces excessive interference to stations operating on adjacent channels because of the production of broad side bands.

The circuit of figure 8–7 may be regarded as being composed of two sources of signal voltage and a common load in series. One source is the r-f generator, the other is the a-f generator, and the load is the resonant tank circuit coupled to the antenna load.

Whenever two or more voltages of different frequencies are

introduced into a circuit having a common nonlinear load impedance (the class-c amplifier), these two voltages combine to produce two additional frequencies. These additional frequencies are called SIDE BANDS or SIDE-BAND FREQUENCIES, as explained in the beginning of this chapter. They are the sum-and-difference frequencies of the r-f and a-f generators.

Thus, in the example under consideration, the r-f generator supplies 100 watts of power at a frequency of 1 mc; and the a-f generator (in the case of 100-percent modulation) supplies 50 watts of a-f power at a frequency of 1 kc. The tank circuit and the antenna coupled to it may be regarded as containing not only the r-f carrier current and its associated field (fig. 8–8) but also two closely associated currents and

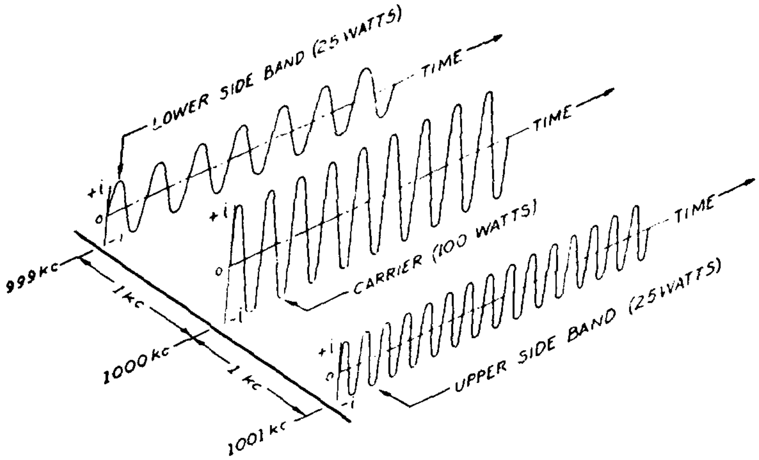


Figure 8–8.—Carrier wave and its side-band frequencies.

their fields—one current having a frequency of 1000+1, or 1001 kc, and the other having a frequency of 1000−1, or 999 kc.

The field radiated by the antenna into space at the speed of light may be regarded as being composed of the carrier and these two side bands. The carrier has a power of 100 watts,

neglecting circuit losses. The 50 watts of audio power are equally distributed in each side band. Thus, the 999-kc side band has a power of 25 watts, and the 1001-kc side band also has a power of 25 watts. When the r-f carrier is 100-percent modulated, one-sixth of the TOTAL power is contained in each side band.

During modulation, the r-f amplifier must handle peak currents that are twice the normal (unmodulated) value. Thus, during modulation an amplifier must be capable of handling up to four times the power it dissipates during steady intervals of unmodulated carrier output. For this reason, in a transmitter that is designed for both continuous-wave (c-w) and radiotelephone service, the modulated amplifier stages are ordinarily reduced in carrier power output for phone operation. Even if the power output is not reduced, the phone signal is weaker because it depends on the side-band power, which cannot exceed one-half of the carrier power.

The example under discussion has involved the simplest form of amplitude modulation. A single tone having sine waveform—like that produced by a tuning fork—constitutes the a-f input. Two side-band components accompany the r-f carrier (fig. 8–8); and when the modulation is 100 percent the power in each side band is one-sixth of the total input power. When the audio signal becomes more complex, the number of different frequency components increases.

For each a-f component a sum-and difference frequency is generated in the output of the transmitter. The power output of the audio modulator is divided equally among the side-band frequencies. Thus, as the number of side-band frequencies increases, the amount of power contained in any one side-band frequency is reduced. For example, for a given audio power input a signal containing speech does not have as much strength as a code signal modulated by a single 1,000-cps tone. For this reason a tone-modulated code signal has a slightly greater distance range than voice modulation, for the same transmitter.

There are various methods of producing an amplitude-

modulated wave, but the most important one is high-level plate modulation. It has been shown that the audio signal is introduced in series with the plate circuit of a class-C amplifier where the plate voltage is at a relatively high level—the highest power level of the entire system. A class-C amplifier using high-level plate modulation is more efficient than a class-B or class-A amplifier that must be used with low-level modulation. Furthermore, class-C amplifiers are more easily adjusted and have proportionately less plate-power loss. For these reasons, high-level plate modulation is widely used.

GRID MODULATION.—Low-level grid modulation requires less bulky equipment than high-level plate modulation, with consequent savings in space, weight, and input power. This type of modulation has many Navy shipboard and aircraft applications. The a-f signal is applied in series with the grid circuit of the r-f power amplifier tube. The a-f signal varies the grid bias, which in turn varies the power output of the r-f amplifier. This variation in power output causes a modulated wave to be radiated. This method is known as GRID-BIAS MODULATION.

A circuit using grid-bias modulation is shown in figure 8-9. A modulation transformer is placed in series with the grid

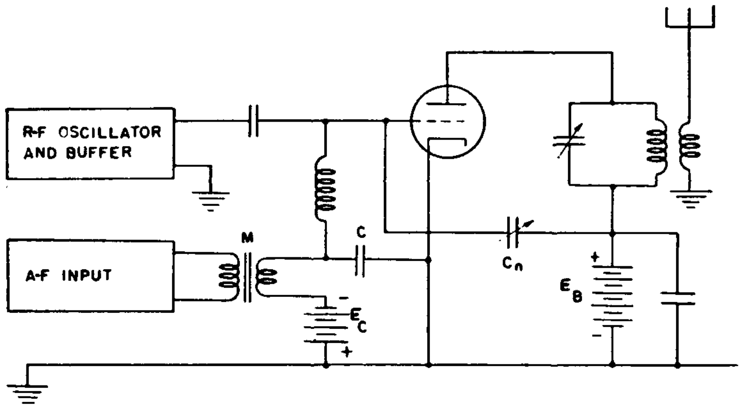


Figure 8-9.—Grid-bias modulation.

return lead of the r-f power amplifier. The a-f voltage from a modulating amplifier adds to or subtracts from the fixed grid-bias voltage and thus controls the output power from the r-f amplifier. The audio modulator tube supplying the modulation transformer, M, must be operated as a class-A amplifier. Varying the grid bias of the r-f stage does not require a great amount of power. It is difficult to achieve any large degree of modulation by this method, and the r-f carrier output power is about one-quarter of that of the plate-modulated transmitter; thus, the intelligibility of the signal is decreased.

TONE TRANSMISSION.—When c-w radiotelegraph signals are being received, the pitch of the sound in the headset depends upon the difference between the incoming signal frequency and the frequency of the heterodyne oscillator—in other words, the beat frequency.

If the frequency of the c-w transmitter varies, the pitch of the received tone will vary. If the drift of the transmitter frequency is very great, the received signal may become inaudible. Under these conditions, the reception of c-w signals becomes very difficult. An obvious remedy is to stabilize the c-w transmitter frequency, but this is not always practicable or possible. In such an event, communication by radiotelegraph may be maintained by using a tone-modulated wave. This method is known as MODULATED-CONTINUOUS WAVE (m-c-w), or TONE, TRANSMISSION and is available for use on most Navy transmitters.

In tone transmission the r-f carrier is modulated at a fixed audio frequency of about 1,000 cps. The output of the transmitter is keyed in the same manner as for c-w transmission. An additional relay reduces the output of the tone generator to zero when the key is up and the r-f output is zero. Because a buzzer or audio oscillator is generally used as the tone source, the amplitude of the modulated r-f output wave is practically constant and the modulation can be 100 percent. Tone modulation has a slightly greater distance range than voice modulation for the same transmitter. However, the range of tone modulation is always less than

that of c-w transmission on the same transmitter. Receiver tuning on tone signals is broader than in straight c-w reception, and code signals are less apt to be lost as a result of transmitter-frequency drift.

FREQUENCY MODULATION

The intelligence to be transmitted may be superimposed on the carrier in the form of changes in the frequency of the carrier. This type of modulation is called FREQUENCY MODULATION and has certain inherent advantages over conventional a-m transmission, particularly when static-free transmission is desired. However, in extensive tests conducted at the Naval Research Lab (NRL), it was found that for general Navy use amplitude modulation was in many ways more desirable than narrow-band frequency modulation. Nevertheless, the Naval Communication System uses a limited number of f-m transmitters and receivers. Aircraft altimeters use frequency modulation, as do some other radar and sonar equipments.

F-M Side Bands

An a-m wave contains one upper and one lower side band for each modulating frequency, and they are in phase with the carrier. An f-m wave may contain more than one pair of side-band frequencies for each modulating frequency, and they are out of phase with the carrier.

As shown in figure 8–10, the f-m wave consists of a carrier wave of frequency f_o and associated side-band frequencies of $f_o \pm f_m$, $f_o \pm 2f_m$, $f_o \pm 3f_m$, and so forth, where f_m is the modulating frequency and f_o is the carrier frequency. Each line on each side of the center line represents a particular component of the f-m wave. The center line represents the carrier. The lines to the right of center represent the upper side-band frequency components, and those to the left of center represent the lower side-band components. The lengths of the lines represent the energy levels of the various components. The horizontal distance between the center line and the last

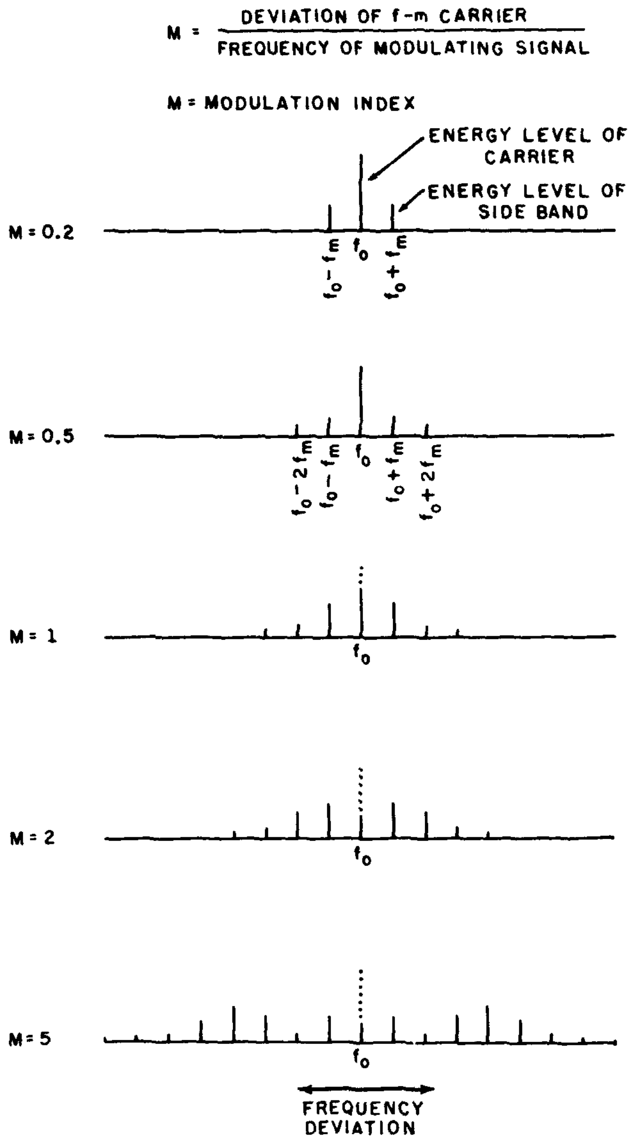


Figure 8–10.—Frequency and energy distribution for 5 values of modulation index of an f-m wave.

significant side band (farthest removed from the center line) is proportional to the deviation frequency of the carrier, which depends in turn on the amplitude of the modulating frequency, f_m. The side-band frequency components are spaced an amount equal to the modulating frequency.

The number of side-band frequencies that contain sufficient energy to be important depends on the frequency deviation imposed on the carrier by the modulating frequency. For example, if the modulating frequency, f_m, causes the carrier to deviate an amount equal to f_m and no more, the first pair of side bands, $f_o \pm f_m$, is the only pair of importance. No additional energy is supplied an f-m wave during modulation. In contrast, additional energy is supplied an a-m wave during modulation. The energy of the f-m wave is redistributed during modulation, as shown in figure 8–10.

The carrier energy is reduced when the modulation index (M) exceeds 0.5. The modulation index is $\frac{f_d}{f_m}$, where f_d is the frequency deviation of the carrier and f_m is the modulating frequency. During modulation, energy is taken from the carrier and redistributed in the side-band components. For the cases $M=2$ and $M=5$, some of the side-band frequency components contain more energy than the carrier. During strong modulating signals the energy level of the carrier approaches zero.

Starting at the carrier and counting the side-band components consecutively in each direction, the upper odd-numbered side-band frequency component is 180° out of phase (the phase relations are not shown in the figure) with its associated lower side-band component. Thus the upper odd-numbered side-band component, f_o+f_m, is 180° out of phase with the odd-numbered lower side-band component, f_o-f_m. All even-numbered side-band frequency components are in phase with each other. Thus f_o+2f_m is in phase with f_o-2f_m. The energy levels of a given pair equally spaced from the carrier are always equal.

As with amplitude modulation, the bandwidth for frequency modulation or phase modulation is determined by the

number of side bands associated with the carrier. An f-m wave with single-tone modulation theoretically has an infinite number of side-band pairs instead of just one pair, as in amplitude modulation. Fortunately, however, only a limited number of side bands contain sufficient energy to be significant. An approximation of the number of significant side-band frequencies may be made by assuming that the important side-band components extend over a frequency range on each side of the carrier by an amount equal to the sum of the modulation frequency and the carrier frequency deviation.

For example, in figure 8–10 the frequency deviation of the carrier, f_o, is assumed to be 50 kc when the modulating frequency, f_m, is equal to 10 kc (M=5) and the bandwidth on each side of the carrier is approximately 50+10, or 60 kc, making a total bandwidth of 60×2, or 120 kc. Because the side-band components are spaced an amount equal to the modulation frequency, the product of the number of significant side-band components and the modulation frequency is equal to the bandwidth. In this example there are 8 significant side-band frequencies on each side of the carrier, as shown in the illustration at the bottom of figure 8–10. Thus the bandwidth is 8×10, or 80 kc, on each side of the carrier, or a total of 2×80, or 160 kc. The bandwidth thus depends upon (1) the modulation frequency and (2) the total frequency deviation of the carrier.

Thus the bandwidth requirement of an f-m system is greater than the frequency deviation of the carrier by an amount equal to at least the modulating frequency. For most f-m systems the bandwidth is greater than that required for a-m systems. F-m transmission is made on higher carrier frequencies (88 to 108 mc for commercial channels) to obtain the necessary number of wide-band channels.

For commercial high-fidelity broadcast transmission, 15 kc is the highest modulation frequency. The maximum frequency deviation of the carrier is limited by the Federal Communications Commission (FCC) to 75 kc on each side of the carrier frequency. The ratio of frequency deviation

to modulation-frequency, or modulation index M, is therefore $\frac{75}{15}=5$. As previously noted for a modulation index of 5, there are 8 important side bands. Because the side bands are spaced 15 kc apart, the bandwidth requirement is 8×15, or 120 kc, on each side of the carrier.

Although the FCC regulation limits the carrier shift to ± 75 kc, some significant side bands may extend beyond this frequency. A guard band of 25 kc on each side of the allowable frequency swing of ± 75 kc is established to take care of most of the significant side bands beyond the established limits.

Naval communications sets do not need high-fidelity response and therefore use modulating frequencies up to only a few thousand cycles. As shown in figure 8–10, when lower frequencies are used to modulate the carrier, the required bandwidth becomes narrower and approaches a value equal to twice the maximum frequency deviation of the carrier. Narrow-band frequency modulation, therefore, requires a bandwidth of approximately twice the maximum frequency deviation. The maximum frequency deviation is about 15 kc; therefore, the bandwidth is approximately 2×15, or 30 kc.

Degree of Modulation

To explain 100-percent modulation in an f-m system, it is desirable to first review the same condition for an a-m wave. As has been stated, 100-percent modulation exists when the amplitude of the carrier varies between zero and twice its normal unmodulated value. There is a corresponding increase in power of 50 percent. The amount of power increase depends upon the degree of modulation; and because the degree of modulation varies, the tubes cannot be operated at maximum efficiency continuously.

In frequency modulation, 100-percent modulation has a different meaning. The a-f signal varies only the frequency of the oscillator. Therefore, the tubes operate at maximum efficiency continuously and the f-m signal has a constant

power input at the transmitting antenna regardless of the degree of modulation. A modulation of 100 percent simply means that the carrier is deviated in frequency by the full permissible amount. For example, an 88-mc f-m station has 100-percent modulation when its audio signal deviates the carrier 75 kc above and 75 kc below the 88-mc value, when this value is assumed to be the maximum permissible frequency swing. For 50-percent modulation, the frequency would be deviated 37.5 kc above and below the resting frequency.

Systems of Frequency Modulation

A successful f-m transmitter must fulfill two important requirements—(1) the frequency deviation must be symmetrical about a fixed frequency, and (2) the deviation must be directly proportional to the amplitude of the modulation and independent of the modulation frequency.

There are several systems of frequency modulation that fulfill these requirements. A MECHANICAL MODULATOR employing a capacitor microphone is the simplest system of frequency modulation, but is seldom used. The two most important systems of frequency modulation are REACTANCE-TUBE and ANGLE, or PHASE-ANGLE, MODULATION. The main difference between these two systems is that in reactance-tube modulation the r-f wave is modulated at its source (the oscillator), while in phase modulation the r-f wave is modulated in some stage following the oscillator. The results of each of these systems are the same—that is, the f-m wave created by either system can be received by the same receiver.

CAPACITOR-MICROPHONE SYSTEM.—The simplest form of frequency modulation is that of a capacitor microphone, which shunts the oscillator-tank circuit, *LC*, as shown in figure 8–11. The capacitor microphone is equivalent to an air-dielectric capacitor, one plate of which forms the diaphragm of the microphone. Sound waves striking the diaphragm compress and release it, thus causing the capacitance

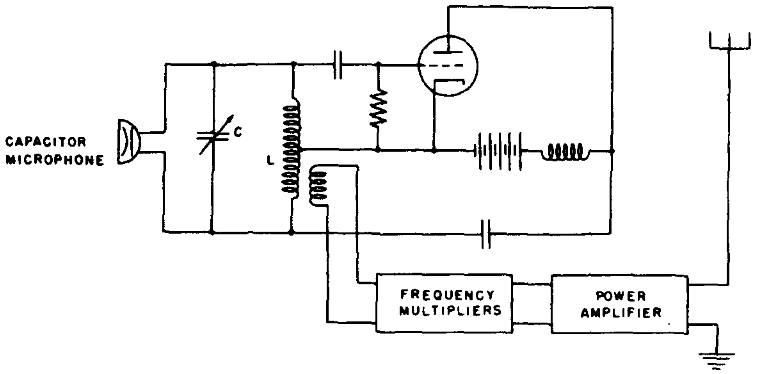


Figure 8–11.—F-m transmitter modulated by a capacitor microphone.

to vary in accordance with the spacing between the plates. This type of transmitter is not practicable (among other reasons, the frequency deviation is very limited), but it is useful in explaining the principles of frequency modulation. The oscillator frequency depends on the inductance and capacitance of the tank circuit, *LC*, and therefore varies in accordance with the changing capacitance of the capacitor microphone.

If the sound waves vibrate the microphone diaphragm at a low frequency, the oscillator frequency is changed only a few times per second. If the sound frequency is higher, the oscillator frequency is changed more times per second. When the sound waves have low amplitude, the extent of the oscillator frequency change from the no-signal, or resting, frequency is small. A loud a-f signal changes the capacitance a greater amount and therefore deviates the oscillator frequency to a greater degree. Thus, the deviation frequency of the oscillator tank depends upon the amplitude of the modulating signal.

In some military systems, in order to prevent interference between adjacent channel f-m transmitters, a bandwidth of 80 kc plus a guard channel of 20 kc is allowed for each transmitter. Thus, the strongest audio signal that can be used for modulating an f-m transmitter is limited to the value

that causes a maximum deviation of 40 kc on each side of the average, or resting, carrier frequency. This allowance makes available a total of 80 kc, known as the CARRIER SWING, over which the frequency of any one transmitter may vary.

To increase the initial deviation frequency of the oscillator (which is greatly restricted in the case of the capacitor-microphone modulator) to a suitable value in the output, a system of frequency multiplication is used. The circuits used to accomplish this frequency multiplication are contained in the block diagram labeled "frequency multipliers" in figure 8–11. One method uses a broadly tuned plate-tank circuit in a class-C amplifier. The tank is tuned to the second harmonic of the grid-input signal and thus builds up a tank current and output signal at double frequency. The output of the first doubler is fed into another similar doubler. Actually several stages may be used. For example, a 5-mc signal fed into a 3-stage amplifier using frequency doubling becomes a 40-mc signal at the output. An initial deviation of 1 kc produced by an audio-modulation signal becomes a frequency deviation of 8 kc at the output of the third doubler stage. Actually, frequency triplers or quadruplers may be used in some systems.

REACTANCE-TUBE SYSTEM.—The reactance-tube system of frequency modulation is shown in figure 8–12. The reactance tube is an electron tube operated so that its reactance varies with the modulation signal and thereby varies the frequency of the oscillator stage. The oscillator,

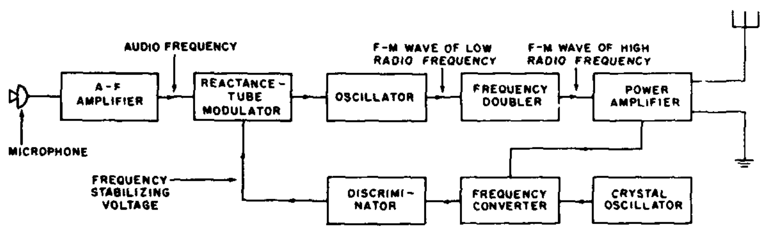


Figure 8–12.—Block diagram of a reactance-tube f-m transmitter.

connected between the reactance tube and the frequency doubler, is self-excited.

In this circuit the reactance tube is connected in parallel with the oscillator tank and functions like a capacitor whose capacitance is varied in accordance with the audio signal, as in the capacitor-microphone system of frequency modulation. The frequency of the oscillator is thus changed, and the resulting f-m signal is passed through a frequency doubler to increase the carrier frequency and the deviation frequency. A power amplifier feeds the final signal to the antenna. The transmitter is kept within its assigned frequency limits by comparing the output of the transmitter with that of a standard crystal-controlled oscillator, and feeding back a suitable correcting voltage from a frequency-converter and discriminator (frequency-detector) stage.

The theory of operation of a reactance-tube circuit may be explained with the aid of figure 8–13. The reactance tube, V1 (fig. 8–13, A), is effectively in shunt with the oscillator tank, *LC*, and the phase-shift circuit, $R_g C1$. The capacitive reactance of the capacitor is large compared with the resistance of the resistor; and the current, i, in this circuit leads the voltage, e_p, across the circuit by approximately 90°. The voltage, e_p, is the alternating component of the plate-to-ground voltage appearing simultaneously across the reactance tube, the phase-shift circuit, and the oscillator tank.

The coupling capacitor, *C*2, has relatively low capacitive reactance to the a-c component of current through it, and at the same time it blocks the d-c plate voltage from the phase-shift circuit and the tank. The reactance tube receives its a-c grid-input voltage, e_g, across R_g. This voltage is the *IR* drop across R_g and is in phase with plate current i_p and grid voltage e_g. This relation is characteristic of amplifier tubes.

Because e_g is in phase with both i and i_p and e_g leads e_p by approximately 90°, both i and i_p lead e_p by approximately 90°. These relations are shown in the vector diagram in figure 8–13, B. Both i and i_p are supplied by the oscillator tank circuit, and because both are leading currents with

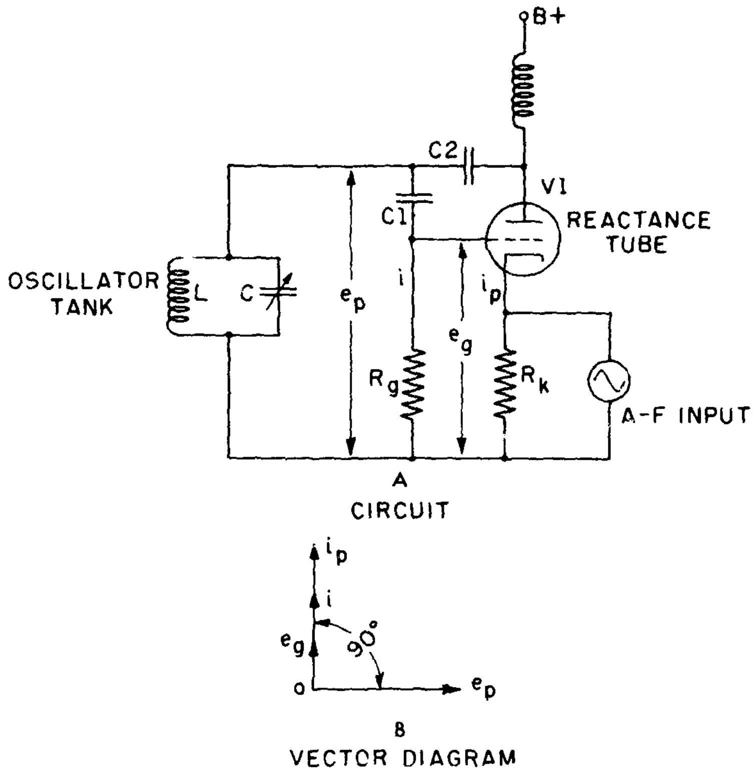


Figure 8–13.—Frequency modulation with a reactance-tube modulator.

respect to the tank voltage, e_p, they act like the current in tank capacitor C. Therefore, the EFFECT of these currents on the frequency of the tank is the same as though additional capacitance were connected in parallel with it.

Consider now the effect of introducing an audio signal across R_k. With zero audio voltage, r-f plate current i_p is a succession of rapid pulses of constant amplitude, and the oscillator tank operates at a constant frequency, called the NO-SIGNAL, or RESTING FREQUENCY. When the audio voltage rises with a polarity that swings the cathode negative

with respect to the grid, the pulses of plate current gradually increase in amplitude. This leading r-f plate current is drawn through the oscillator tank and is equivalent to an increasing value of tank capacitance. Thus, the oscillator frequency is lowered.

Conversely, when the audio signal swings the grid of the reactance tube negative with respect to the cathode, the r-f plate-current pulses gradually decrease in amplitude and the oscillator frequency increases.

The frequency of the a-f signal determines the number of times per second that the oscillator-tank frequency changes. On the other hand, the amplitude of the a-f signal determines the extent of the oscillatory-frequency change—that is, the deviation frequency Thus, the reactance tube with its audio-signal input produces an f-m output having the same characteristics as that of the capacitor-microphone modulator.

PHASE MODULATION.—Any process that changes the instantaneous frequency of the r-f energy already generated at a constant frequency is referred to as angle, or phase-angle, modulation. All radio modulating processes are based on changing the r-f carrier wave in some respect. The variation normally is directly proportional to the instantaneous value of the modulating voltage. When the instantaneous frequency of the carrier is varied in a direct relation to the modulating wave, the result is frequency modulation. If the instantaneous phase of the carrier is varied by an electrical angle directly proportional to the instantaneous modulating voltage, phase modulation is obtained. Varying the carrier frequency also changes the instantaneous phase relation of the carrier frequency to its own fixed unmodulated state. Likewise, varying the carrier phase changes the carrier frequency.

Thus, frequency modulation and phase modulation are basically the same. In fact, frequency modulation is equivalent to phase modulation in which the phase-angle variation is inversely proportional to the modulation frequency. Similarly, phase modulation is equivalent to frequency

modulation when it has preemphasis (the signal strength is increased as the audio frequency becomes higher) over the entire range of modulation frequency. An f-m signal is one in which the carrier deviation from the resisting frequency is proportional to the amplitude of the modulating signal and is independent of the modulating frequency. A pure phase-modulated signal is one in which the carrier deviation is proportional to both the amplitude and frequency of the modulating signal.

In all cases the carrier variation occurs at a rate of change equal to the frequency of the modulating wave. For example, a 1,000-cps tone changes the carrier frequency plus and minus 75 kc 1,000 times per second in a broadcast f-m system at maximum modulation.

A phase-modulation system is shown in the block diagram in figure 8–14. The transmitter oscillator is maintained at

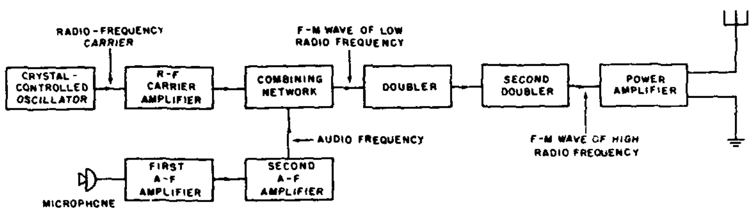


Figure 8–14.—Block diagram of a phase-modulated f-m transmitter.

a constant frequency by means of a quartz crystal. This constant-frequency signal passes through an amplifier that increases the amplitude, or energy level, of the wave. The audio signal is applied to the r-f carrier by means of a combining network. The output of the combining network is fed into a series of class-C amplifiers, the plate circuits of which are tuned to a multiple (doubling is indicated in this figure) of the frequency of the grid input. The output of these frequency multipliers is fed to a power amplifier that couples the f-m signal to the antenna.

A diagram of the combining network in which the phase shift is accomplished is shown in figure 8–15, A. The r-f and a-f voltages are applied across the grid-input voltage

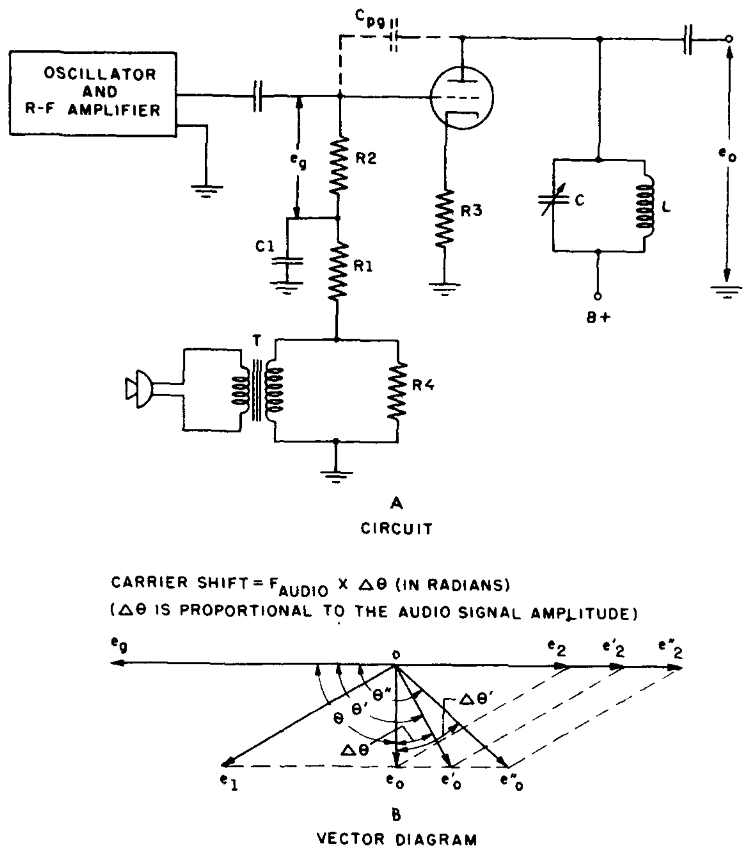


Figure 8-15.—Phase-modulated network.

divider, which consists of $R2$, $R1$, and $R4$. The triode plate load is a broadly tuned L-C tank. The r-f signal of constant frequency and amplitude appears across $R2$ as e_g. As the instantaneous value of the audio signal varies through each audio cycle, the triode bias is increased and decreased at the a-f ratio because of the a-f voltage that appears across resistors $R4$ and $R1$. Consequently, the triode gain is varied in accordance with the a-f signal.

Now consider how this varying gain is translated into phase shift. The instantaneous plate-load voltage, e_o (fig. 8–15, B), is the resultant of two r-f voltages in the triode plate circuit. These two r-f voltages are (1) that portion of the grid-input signal that is coupled to the plate circuit by means of the grid-plate capacitance of the triode (this voltage is designated e_1) and (2) the grid-input signal amplified in the triode plate by normal amplifier action (this voltage is designated e_2).

The resultant of e_1 and e_2 is indicated in the vector diagram as e_o. In the same diagram, e_g is the grid-input r-f signal. The triode-amplifier voltage, e_2, is relatively low because of inverse feedback obtained by omitting the usual cathode-bypass capacitor across $R3$. Voltage e_1 leads e_g and is of less amplitude than e_g because the interelectrode capacitive reactance acts in series with the plate load and causes a leading current to flow through it. The plate load is resonant at the oscillator frequency and hence acts as a pure resistance. Thus e_1 across the load is in phase with the leading current through the load and leads e_g by some angle less than 90° depending on the magnitude of the interelectrode capacitive reactance and the plate-load resistance.

At the time the audio frequency swings the triode bias to the MAXIMUM the triode gain is a MINIMUM, and e_2 is relatively small. For this condition e_o leads e_g by angle θ. The amplified voltage, e_2', represents the condition existing when the a-f signal swings the triode bias to a minimum, and the tube gain is higher than before. Therefore, voltage e_2' is larger and combines with e_1 to produce the resultant plate-load voltage, e_o', which leads e_g by angle $\theta+\Delta\theta$. The resultant voltage, e_o', undergoes a change in phase angle, $\Delta\theta$, with respect to e_g in accordance with the change in triode bias, gain, and a-f instantaneous values.

The difference in the angles θ and θ'—that is, $\Delta\theta$—is the change in the PHASE-SHIFT ANGLE of e_o, and is a factor of carrier swing. When voltage variations, e_o, e_o', and so forth, are applied to a tuned circuit, a smooth wave having both positive and negative alternations is formed as the result of

the flywheel effect in the tank circuit. This wave has the same varying time internal between positive peaks as the applied voltage variations, and therefore its frequency is shifted in accordance with the audio-modulation signal during the time the phase angle is changing.

In a phase-modulation system such as the one being discussed the carrier shift is proportional to the product of the audio frequency and the phase-shift angle. It is therefore necessary to introduce an action that will prevent a signal that changes in audio frequency, but remains at constant amplitude, from influencing carrier swing. Otherwise low-frequency audio components would be underemphasized at the receiver. Only the AMPLITUDE of the modulating signal and NOT its instantaneous frequency should influence the extent of the frequency swing of the carrier. This action may be accomplished by introducing a preemphasis circuit such as the one composed of $R1C1$ in figure 8–15, A.

When the audio signal of constant amplitude is decreased *from a high frequency to a lower frequency*, the audio voltage across $C1$ is increased in amplitude. Therefore, the tube-bias swing and the change in tube gain cause the normal amplifier component of output voltage to be increased from e_2e_2' to e_2e_2''. Carrier shift is proportional to the product of the audio frequency, f_{audio}, and the phase-shift angle, $\Delta\theta$, as indicated in figure 8–15, B. Then an increase in the phase-shift angle from $\Delta\theta$ to $\Delta\theta'$ compensates for the decrease in audio frequency, f_{audio}. Thus, the product of f_{audio} and $\Delta\theta$ remains constant, and the carrier swing is now independent of the audio frequency. The output f-m signal from the phase-modulated transmitter is therefore similar to that of a frequency-modulated transmitter. Thus, an f-m receiver performs equally well on either output.

DEMODULATION OF F-M WAVES

DEMODULATION, or DETECTION, is the process of recovering the intelligence from a modulated wave. When a radio carrier wave is amplitude-modulated, the intelligence is imposed on the carrier in the form of amplitude variations

of the carrier. The demodulator of an amplitude-modulated (a-m) wave produces currents or voltages that vary with the amplitude of the wave. Likewise, the frequency-modulation (f-m) detector and the phase-modulation (p-m) detector change the frequency variations of an f-m wave, and the equivalent phase variations of a p-m wave, into currents or voltages that vary in amplitude with the frequency of phase changes of the carrier.

The instantaneous amplitude, e_o, of the carrier may be represented as

$$e_o = E_o \sin (2\pi f_o t + \theta),$$

where E_o is the maximum amplitude of the original carrier, f_o the frequency of the carrier, and θ the phase angle (for a-m signals, θ may be considered as zero). One or more of the independent variables (those on the right-hand side of the equation) may be made to vary in accordance with the modulating signal to produce a variation in e_o. However, the general practice is to vary only one of the values—E_o (for a-m), f_o (for f-m), or θ (for p-m)—and to prevent any variation in the others.

The detector in the receiver must therefore be designed so that it will be sensitive to the type of modulation used at the transmitter, and insensitive to any other.

Most Navy equipment is designed for amplitude modulation. A clear understanding of the mechanism of a-m detection is therefore very important.

A-m modulators and demodulators are nonlinear devices. A NONLINEAR DEVICE is one whose current-voltage relation is not a straight line. Because the ratio of current to voltage is not constant, the device has a nonlinear impedance. When an a-m wave is impressed on a nonlinear impedance—for example, one of the electron-tube detectors to be considered later—the average output current is the difference between each successive positive and negative swing of the output signal current, as shown in figure 8–16. The average output (signal component) follows the envelope of the incoming

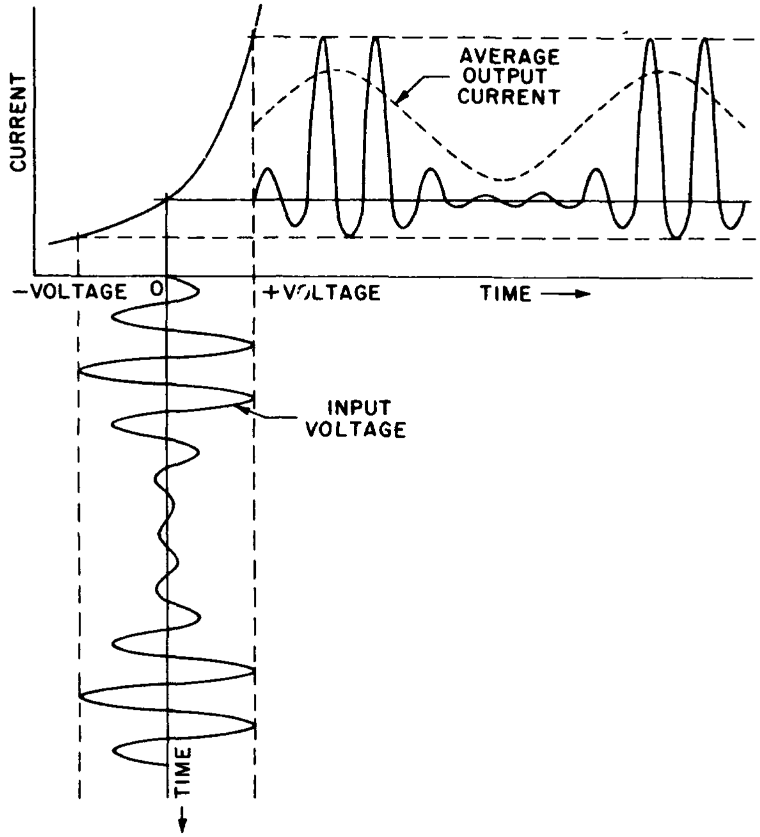


Figure 8-16.—Results of impressing an a-m wave on a nonlinear device.

modulated wave more or less closely, depending on the shape of the nonlinear curve. Because the envelope of the incoming a-m wave contains the desired audio frequency, a nonlinear device demodulates the a-m wave.

For an understanding of the differences in the output frequencies of the various detectors it is necessary to examine the frequencies involved in both modulation and demodulation.

Comparison of Amplitude Modulation and Demodulation

If at the transmitter an r-f carrier and a single-frequency audio-modulating signal of sine waveform are impressed on a LINEAR device, the output waveform from the linear device will contain the same r-f and a-f signal frequencies. The tuned r-f amplifiers in the transmitter will amplify the r-f carrier, but will eliminate for all practical purposes the a-f component. Under these circumstances only the carrier will be radiated, and it will be ineffective in "carrying" the intelligence component.

A very different result is obtained if an r-f carrier and a single-frequency audio-modulating signal of sine waveform are impressed on a NONLINEAR device. In this instance distortion is introduced and, as a result, additional frequencies are produced. In addition to the original frequencies, sum-and-difference frequencies are generated, and a zero-frequency, or d-c, component, is added. The tuned circuits at the transmitter now respond to the carrier and the upper and lower side bands; but, as before, the a-f modulating signal is discriminated against. However, this a-f component is replaced, or generated, by the demodulator in the receiver.

In the receiver the carrier and the two side bands are impressed on a second nonlinear device called the DEMODULATOR. If the demodulator has an IDEAL nonlinear curve it will distort the incoming waveform (the positive halves of the cycle will be different from the negative halves). Therefore, in addition to the r-f carrier and the two side bands, the SIGNAL FREQUENCY (which is the difference between the upper side band and the carrier or the difference between the carrier and the lower side band) and a zero-frequency (or d-c component) will be produced. As shown in chapter 12, the d-c component may be used for automatic volume control.

If the demodulator used in the receiver does not have an ideal nonlinear curve, but has a PRACTICAL realizable curve such as the square-law curve, additional frequencies will be produced. These frequencies will be harmonics of all fre-

quencies present in the input. They are produced because input voltages having larger amplitudes are distorted differently from input voltages having smaller amplitudes. The r-f harmonics may be filtered in the output of the demodulator, but the a-f harmonics are not easily eliminated.

Thus, modulation and demodulation are essentially the same in that the waveform is distorted in each case and new frequencies are produced.

Types of A-M Detectors

Detectors are classified according to the shape of their current-voltage (characteristic) curve. If the curve is smooth, as in figure 8–16, the detector is called a SQUARE-LAW DETECTOR. It is called a square-law detector because, for a first approximation, the output voltage is proportional to the square of the effective input voltage.

If the current-voltage curve of the detector is shaped like an obtuse angle, as in figure 8–17, A, the curve is still nonlinear because of the abrupt change in shape at the knee. Because the detector action takes place on the linear portions of the curve on both sides of the knee, this type of detector is called a LINEAR DETECTOR. It is called a linear detector merely to distinguish it from a square-law detector. Both the square-law detector and the linear detector are actually NONLINEAR devices.

The rectified output voltage of a linear detector is proportional to the input voltage. The output of a square-law detector is proportional to the square of the input voltage.

Detectors are also described as power detectors or as weak-signal detectors. If a detector is to handle r-f carrier voltages having amplitudes greater than approximately 1 volt it is called a POWER DETECTOR. If the input signal strength is less than this amount the detector is called a WEAK-SIGNAL DETECTOR. Thus, the approximate value of 1 volt is the dividing line between the two detectors. Because a linear detector cannot be obtained with a sharp discontinuity at the knee, weak signals are always detected on a curved portion of the characteristic, as shown in figure 8–17,

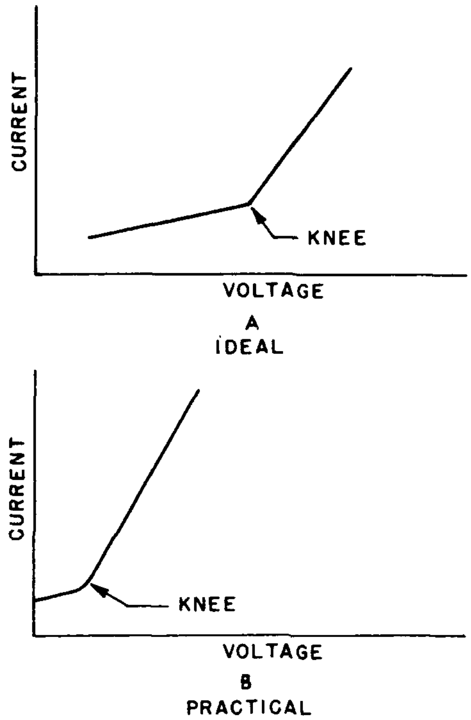


Figure 8–17.—Linear detectors.

B. Thus, weak-signal detectors are always of the square-law type. Power detectors may be either linear or square-law, depending on the application.

DIODE DETECTOR.—The diode detector (fig. 8–18, A) is one of the simplest and most widely used detectors, and has nearly an ideal resistance characteristic. Diodes have a point of sharp discontinuity between the conducting (forward) and nonconducting (reverse) directions and therefore make good detectors.

The slight bend in the lower portion of the i_p-e_p characteristic curve results from contact potential. CONTACT POTENTIAL is the potential difference existing between the

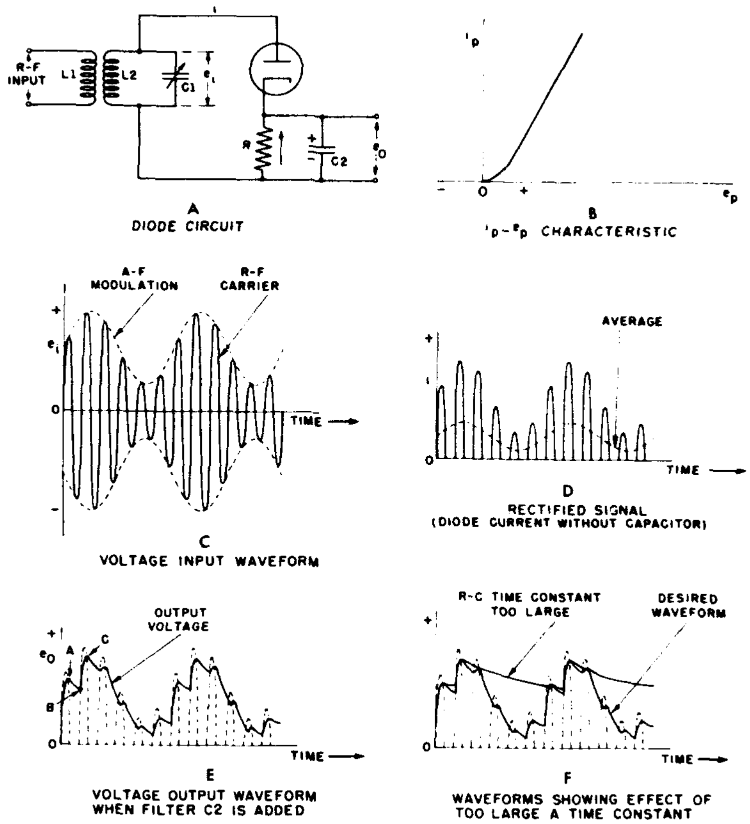


Figure 8–18.—Diode detector and waveforms.

surfaces of metals of different electron affinities that are in direct contact or are connected by means of an external circuit. This is true of the metal elements of electron tubes; and the voltages developed may be of the order of 1 volt or more. In the case of the diode, contact potential keeps the plate current from decreasing to zero when the plate voltage approaches zero. The current, however, is very slight, and the characteristic curve is generally shown as zero when the applied voltage is zero. Nevertheless, on low signal voltage,

plate current does not increase as rapidly as it does after the contact potential has been exceeded.

Because the diode characteristic is nearly straight on both sides of the knee, the diode detector is a linear detector. However, with weak signals, the output of the detector follows the square law because weak signals force the operation to take place on the lower, curved portion of the characteristic curve (fig. 8–18, B). Because the diode detector normally handles large input signals with minimum distortion, it is classified as a power detector.

The modulated signal voltage (fig. 8–18, C) is developed across the tuned circuit, $L2C1$, of the detector stage. Signal current flows through the diode only when the plate is positive with respect to the cathode—that is, only on the positive half cycles of the r-f voltage wave.

The rectified signal flowing through the diode (fig. 8–18, D) actually consists of a series of r-f pulses and not a smooth outline or envelope. The average of these pulses, with little or no filtering, does increase and decrease at the a-f rate, as shown by the dotted line. Therefore, there is an audio voltage output even if no filtering is employed. However, some stray capacitance exists, and consequently some r-f filtering takes place.

If a capacitor ($C2$ in fig. 8–18, A) of the proper size is used as a filter, the output voltage of the detector is increased and more nearly follows the envelope. On the first quarter cycle of applied r-f voltage, $C2$ charges up to nearly the peak value of the r-f voltage (point A in fig. 8–18, E). The small voltage drop in the tube prevents $C2$ from charging up completely. Then as the applied r-f voltage falls below its peak value, some of the charge on $C2$ leaks through R, and the voltage across R drops only a slight amount, to point B. When the r-f voltage applied to the plate on the next cycle exceeds the potential at which the capacitor holds the cathode (point B), diode current again flows and the capacitor charges up to almost the peak value of the second positive half cycle at point C.

Thus the voltage across the capacitor follows very nearly

the peak value of the applied r-f voltage and reproduces the a-f modulation. The detector output, after rectification and filtering, is a d-c voltage that varies at an audio rate, as shown by the solid line in figure 8–18, E. The curve of the output voltage across the capacitor is shown somewhat jagged. Actually, the r-f component of this voltage is negligible and, after amplification, the speech or music originating at the transmitter is faithfully reproduced.

The correct choice of R and $C2$ (fig. 8–18, A) in the diode-detector circuit is very important if maximum sensitivity and fidelity are to be obtained. The load resistor, R, and the plate resistance of the diode act as a voltage divider to the received signal. Therefore, the load resistance should be high compared with the plate resistance of the diode so that maximum output voltage will be obtained. The value of $C2$ should be such that the RC time constant is long compared with the time of one r-f cycle. This is necessary because the capacitor must maintain the voltage across the load resistor during the time when there is no plate current. Also, the RC time constant must be short compared with with the time of one a-f cycle in order that the capacitor voltage can follow the modulation envelope.

The values of R and $C2$ therefore place a limit on the highest modulation (audio) frequency that can be detected. Figure 8–18, F, shows the type of distortion that occurs when the RC time constant is too large. At the higher modulation frequencies the capacitor does not discharge as rapidly as required, and negative peak clipping of the audio signal results.

The efficiency of rectification in a diode is the ratio of the peak voltage appearing across the load to the peak input signal voltage. The efficiency increases with the size of R compared with the diode plate resistance, because R and the diode are in series across the input circuit and their voltages divide in proportion to their resistance. With audio frequencies, a large value of R may be used (of the order of 100,000 ohms), and consequently the efficiency is relatively high (95%). When high modulation frequencies, such as

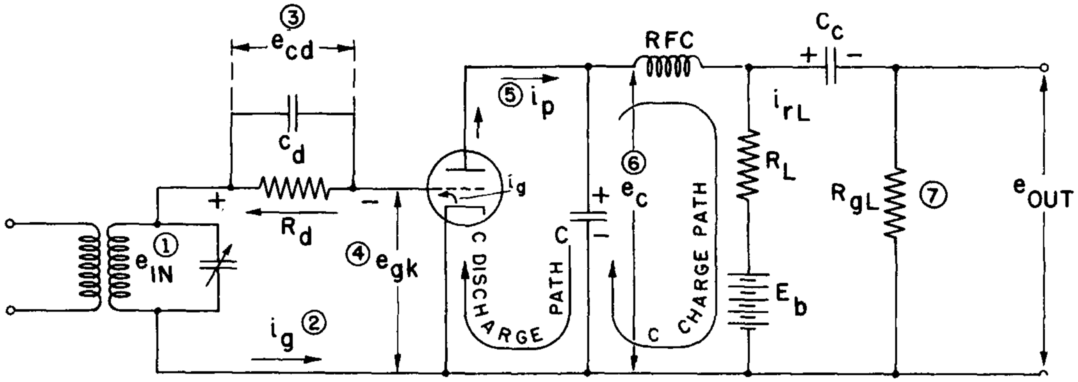
those used in television, are necessary the value of R must be reduced to keep the RC time constant low enough to follow the envelope. Consequently the efficiency is reduced.

The diode detector can handle large signals without overloading, and it can provide automatic-volume-control voltage without extra tubes or special circuits. However, it has the disadvantage of drawing power from the input tuned circuit because the diode and its load form a low-impedance shunt across the circuit.

Consequently, the circuit Q, the sensitivity, and the selectivity are reduced. The interelectrode capacitance of the diode detector limits its usefulness at high carrier frequencies, and the bend in the lower portion of the current-voltage characteristic indicates that it distorts on weak signals. Therefore considerable amplification is needed before detection.

GRID-LEAK DETECTOR.—The grid-leak detector functions like a diode detector combined with a triode amplifier. It is convenient to consider detection and amplification as two separate functions. In figure 8–19, A, the grid functions as the diode plate. The values of C_d and R_d must be so chosen that C_d charges during the positive peaks of the incoming signal and discharges during the negative peaks. The time constant of R_dC_d should be long with respect to the r-f cycle and short with respect to the a-f cycle.

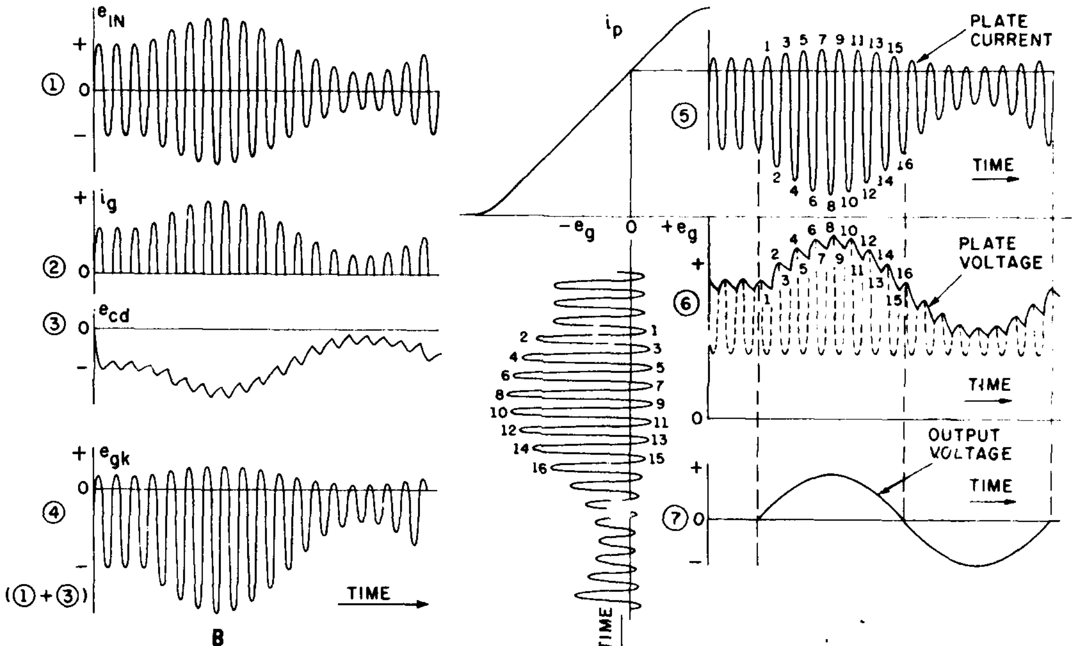
An approximate analysis of the waveforms existing in the diode (grid) circuit is shown in figure 8–19, B. Part (1) shows the input waveform which is also the waveform in the input tuned circuit. Because r-f current i_g flows in only one direction in the grid circuit, part (2) shows a rectified current waveform in this circuit. Part (3) shows the waveform developed across C_d. This audio waveform is produced in the same way as the audio waveform in the diode detector. However, the waveform shown in part (3) is not the output voltage. In the grid-leak detector the waveform produced across C_d is combined in series with the r-f waveform in the tuned circuit to produce the grid-to-cathode waveform shown in part (4).



A

GRID-LEAK DETECTOR CIRCUIT

Figure 8-19.—Grid-leak detector and waveforms.



B
WAVEFORMS IN THE GRID CIRCUIT

C
WAVEFORMS IN THE PLATE CIRCUIT

Figure 8-19.—Grid-leak detector and waveforms—Continued

An approximate analysis of the waveforms existing in the triode plate circuit is shown in figure 8–19, C. Part ⑤ is the plate-current waveform, and part ⑥ is the plate-voltage waveform.

Capacitor C discharges on the positive half cycles of grid input voltage (points 1, 3, 5, 7, 9, 11, 13, and so forth). The discharge path is clockwise through the circuit including the tube and capacitor. The time constant of the discharge path is the product of the effective tube resistance and the capacitance of capacitor C, and this time constant is short because the effective tube resistance is low. The increase in plate current is supplied by the capacitor rather than the B supply, thus preventing any further increase in current through the r-f choke and plate load resistor R_L. Therefore, any further change in plate and capacitor voltage is limited.

Capacitor C charges up as plate voltage rises on the negative half cycles of r-f grid input voltage (fig. 8–19, C, points 2, 4, 6, 8, 10, 12, 14, and so forth). The charging path is clockwise through the circuit containing the capacitor, r-f choke, load resistor R_L, and the B supply. The rise in plate voltage is limited by the capacitor charging current which flows through the r-f choke and through R_L. The plate current decrease is approximately equal to the capacitor charging current; thus the total current through the r-f choke and R_L remains nearly constant, and the plate and capacitor voltage rise is checked.

Positive grid swings cause sufficient grid current flow to produce grid-leak bias. Low plate voltage limits the plate current on no signal in the absence of grid bias. Thus, the amplitude of the input signal is limited, since with low plate voltage the cutoff bias is low, and that portion of the input signal that drives the grid voltage below cutoff is lost. The waveform of the voltage across capacitor C is shown by the solid line in part ⑥ of figure 8–19, C. Part ⑦ shows the output-voltage waveform. This waveform is the difference between the voltage across capacitor C and the d-c voltage across coupling capacitor C_c.

Because the operation of the grid-leak detector depends on a certain amount of grid-current flow, a loading effect is produced which lowers the selectivity of the input circuit. However, the sensitivity of the grid-leak detector is moderately high on low-amplitude signals.

PLATE DETECTOR.—In a GRID-LEAK DETECTOR the incoming r-f signal is detected in the grid circuit and the resultant a-f signal is amplified in the plate circuit. In a PLATE DETECTOR, the r-f signal is first amplified in the plate circuit, and then it is detected in the same circuit.

A plate detector circuit is shown in figure 8–20, A. The cathode bias resistor, $R1$, is chosen so that the grid bias is approximately at cutoff during the time that an input signal of proper strength is applied. Plate current then flows only on the positive swings of grid voltage, during which time average plate current increases. The peak value of the a-c input signal is limited to slightly less than the cutoff bias to prevent driving the grid voltage positive on the positive half cycles of the input signal. Thus, no grid current flows at any time in the input cycle, and the detector does not load the input tuned circuit, $LC1$.

Cathode bypass capacitor $C2$ is large enough to hold the voltage across $R1$ steady at the lowest audio frequency to be detected in the plate circuit. $C3$ is the demodulation capacitor across which the a-f component is developed. $R2$ is the plate load resistor. The r-f choke blocks the r-f component from the output. $R2C3$ has a long time constant with respect to the time for one r-f cycle so that $C3$ resists any voltage change which occurs at the r-f rate. $R2C3$ has a short time constant with respect to the time for one a-f cycle so that the capacitor is capable of charging and discharging at the audio rate.

The action of the plate detector may be demonstrated by the use of the i_p-e_g curve in figure 8–20, B. On the positive half cycle of r-f input signal (point 1) the plate voltage falls below the B supply because of the increased drop across $R2$ and the r-f choke. Capacitor $C3$ discharges. The discharge current flows clockwise through the circuit including

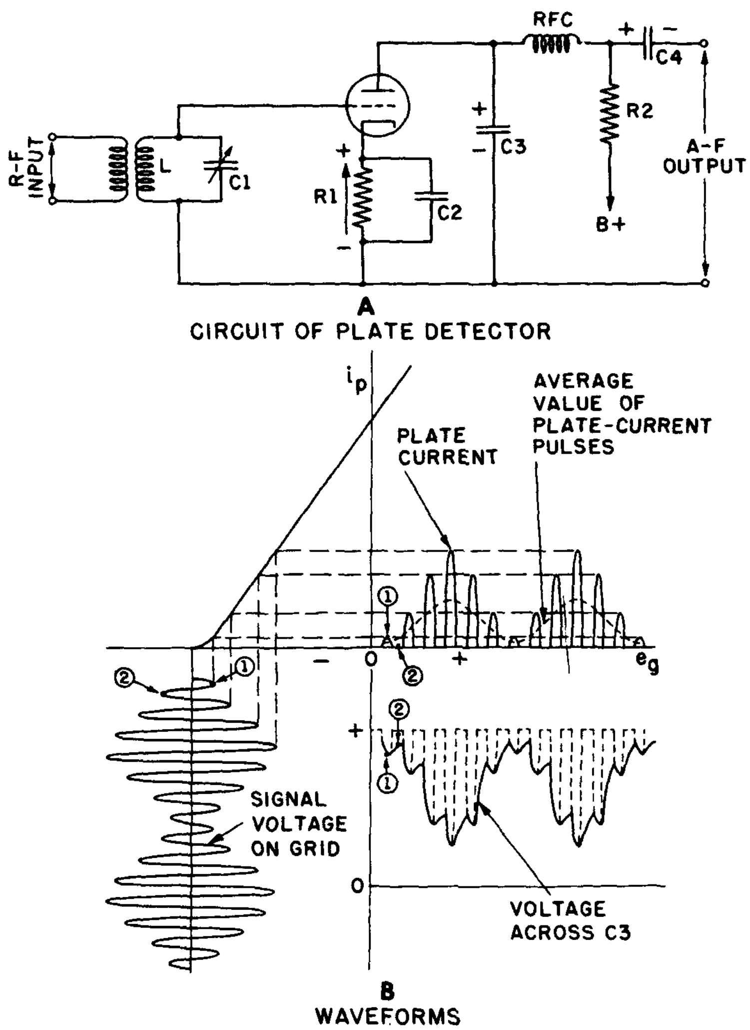


Figure 8-20.—Plate detector and waveforms.

the tube and $C3$. Plate current is supplied by $C3$ rather than the B supply. The drop across $R2$ and the r-f choke is limited, and the decrease in plate voltage is slight.

On the negative half cycle of r-f input signal (point 2) plate current is cut off and plate voltage rises. Capacitor $C3$ charges. The charging current flows clockwise around the circuit including the r-f choke, $R2$, and the B supply. The drop across $R2$ and the r-f choke contributed by the charging current of $C3$ checks the rise in plate voltage.

Thus, $C3$ resists voltage change at the r-f rate. Because $C3R2$ has a short time constant with respect to the lowest a-f signal, the voltage across $C3$ varies at the a-f rate.

The plate detector has excellent selectivity. Its sensitivity (ratio of a-f output to r-f input) is also greater than that of the diode detector. However, it is inferior to the diode detector in that it is unable to handle strong signals without overloading. Another disadvantage is that the operating bias will vary with the strength of the incoming signal and thus cause distortion unless a means is provided to maintain the signal input at a constant level. Thus, a-v-c or manual r-f gain control circuits usually precede the detector.

REGENERATIVE DETECTOR.—When high sensitivity and selectivity are the most important factors to be considered, a regenerative detector may be used. However, the linearity as well as the ability to handle strong signals without overloading is very poor.

The process of feeding some of the output voltage of an electron-tube circuit back into the input circuit so that it adds to or reinforces (is in phase with) the input voltage is known as REGENARATION OR POSITIVE FEEDBACK. The use of regeneration in a circuit greatly increases the amplification of the circuit because the output voltage fed back to the input circuit adds to the original input voltage, thus increasing the total voltage to be amplified by the tube.

A grid-leak detector may be modified to operate as a regenerative detector, as indicated in figure 8–21. Because an amplified r-f component is present in the plate circuit of the grid-leak detector, regeneration can be obtained by con-

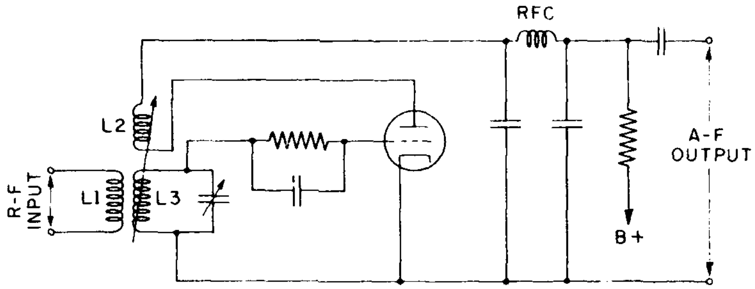


Figure 8-21.—Regenerative detector.

necting a coil, *L2*, known as a TICKLER COIL, in series with the plate circuit so that it is inductively coupled to the grid coil, *L3*.

With an r-f signal across *L3* an r-f component of plate current flows through *L2*. *L2* is connected so that the voltage it induces in *L3* is in phase with the incoming signal voltage applied to the grid. Thus, the voltage gain of the stage is increased.

It is important that the voltage fed back by the tickler coil be in phase with the incoming signal voltage. Otherwise the feedback will be degenerative and the amplification will be reduced. Furthermore, if the coupling between *L2* and *L3* is too great, oscillation will take place. For receiving code signals, oscillation is desirable in order to produce an audible beat tone. However, it is not desirable for voice or music reception because of the objectionable squeal from the beat tone. The regenerative detector is the most sensitive triode detector circuit possible when it is operated just below the point of oscillation.

A comparison of the sensitivity (ratio of a-f output to r-f input), linearity (accuracy of reproduction of the waveform of the modulation component), selectivity (ability to separate signals), and relative ability of the various detectors to handle large signals without overloading is given in table 2.

TABLE 2.—*Comparison of a-m detectors*

Detector type	Sensitivity	Linearity	Selectivity	Ability to handle strong signals without overloading
Diode	Low	Good	Poor	High.
Grid-leak	High	Poor	do	Limited.
Plate	Medium	Fair	Excellent	Medium.
Regenerative	Very high	Poor	do	Poor.

HETERODYNE DETECTOR.—The process of combining two frequencies to obtain the difference frequency is called HETERODYNING. The heterodyne principle has a number of important Navy applications. It is used in heterodyne code reception to change continuous-wave (c-w) telegraph signals to an audio frequency. It is widely used in superheterodyne receivers to change the carrier frequency to the fixed intermediate frequency. Heterodyne action is also employed to separate frequencies that differ from one another by small amounts.

If two a-c signals of different frequencies are combined, or mixed, in a suitable circuit, a third signal, called a BEAT FREQUENCY, will be produced. The frequency of the beat is equal to the difference between the frequencies that are mixed to produce it.

Thus, if two a-f voltages having frequencies of 500 and 600 cps are properly mixed, a beat frequency of 100 cps will be produced by a suitable reproducer.

If two r-f signals differing in frequency by an audible frequency are mixed, an a-f voltage will be produced. Thus, if two r-f voltages having frequencies of 500 and 501 kc are properly mixed, a beat frequency of 1 kc will be produced.

An important application of this principle is in the detection of c-w signals. In c-w transmission the unmodulated carrier is keyed, or interrupted, according to the code that is being transmitted. Because no modulation component is

present in the transmitted energy, the detectors previously considered are not effective in detecting the signal. At best, there may be some noise or sound variation in the reproducer at the beginning and end of each interruption, but reception will be unsatisfactory. In order to receive c-w signals from a radiotelegraph transmitter it is therefore necessary to mix with the incoming signal another locally generated signal that differs in frequency from the incoming signal by some frequency in the audible range.

One method of producing an audible beat tone is by the use of an oscillating regenerative-detector circuit. If the regeneration, or positive feedback, in a regenerative detector is increased beyond a certain critical point, the circuit will oscillate at a frequency approximately equal to the frequency of the tuned circuit. Thus, if the regenerative detector is made an oscillating detector, and is tuned so that the frequency it generates differs from the incoming r-f signal frequency by an audible amount, it is possible to detect unmodulated r-f signals. This process is known as heterodyning, and an oscillating detector is called a HETERODYNE DETECTOR. A brief analysis of the heterodyne principle is included under superheterodyne receivers in chapter 12.

Many communication receivers are equipped to receive both a-m signals and c-w signals. These receivers are commonly designed with a local oscillator, called the BEAT-FREQUENCY OSCILLATOR (BFO), coupled to the plate circuit of the diode detector. The beat-frequency oscillator is then tuned to a frequency that differs from the intermediate-frequency by the desired audio frequency. In t-r-f receivers, the local oscillator must differ from the incoming signal by the desired audio frequency.

The first detector, or frequency converter, of superheterodyne receivers is a heterodyne detector. In this instance there is a modulation component, but it is desirable to reduce the carrier frequency to a new, lower frequency, called the INTERMEDIATE FREQUENCY. Therefore, the incoming signal is mixed with a locally generated signal to produce a fre-

quency that is the difference between the two signals. The modulation envelope is not appreciably affected by the heterodyne action.

DEMODULATION OF F-M WAVES

In f-m transmission the intelligence to be transmitted causes a variation in the instantaneous frequency of the carrier either above or below the center, or resting, frequency. The detecting device must therefore be so constructed that its output will vary linearly according to the instantaneous frequency of the incoming signal. Also, the detecting device must be insensitive to amplitude variation produced by interference or by receiver nonlinearities; thus a special limiting device, called a LIMITER, must precede the f-m detector.

Types of F-M Detectors

A number of f-m detectors might be used. Each has certain inherent advantages and disadvantages. Two of the most common are the DISCRIMINATOR and the RATIO DETECTOR. There are also various types of LOCKED-OSCILLATOR f-m detectors. The simplest type of detector is the SLOPE DETECTOR. Although it is rarely used, this type of f-m detector will be considered first, because of its simplicity.

SLOPE DETECTOR.—Even an a-m receiver may give a distorted reproduction of an f-m signal under certain conditions of operation. When the carrier frequency of the f-m signal falls on the sloping side of the r-f response curve in an a-m receiver, the frequency variations of the carrier signal are converted into equivalent amplitude variations. This conversion results from the unequal response above and below the carrier center frequency (point B), as shown in figure 8–22.

Thus, when the incoming f-m signal is less than the center frequency—for example, at point A, which is the minimum value—the output voltage is at a minimum in the negative direction. When the incoming signal swings to point C (the maximum value), the output voltage is maximum in the posi-

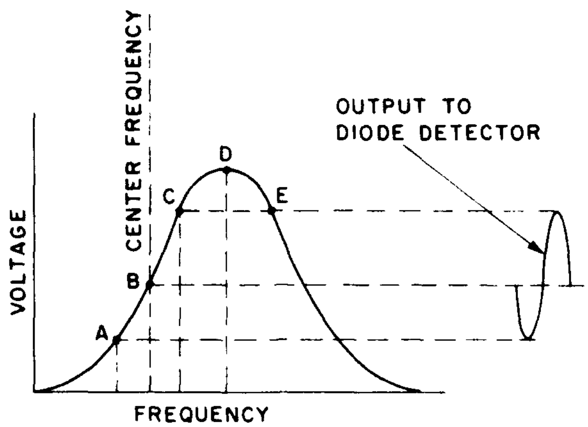


Figure 8–22.—Slope detection.

tive direction. The resultant a-m signal may be coupled to the regular a-m detector where the original audio voltage is recovered.

The obvious disadvantage of this type of detection is the nonlinearity of the response curve. At best, the most linear portion of the curve has a limited frequency range. Consequently the undistorted output voltage is low.

If in figure 8–22 the center frequency falls at point D and the maximum frequency swings are between points C and E, there is no effective output signal voltage because the curve is relatively flat.

Discriminator.—One form of discriminator is shown in figure 8–23. It must be preceded by one or more limiter stages (discussed in chapter 12) because this type of f-m detector is sensitive to both amplitude changes and frequency changes. Basically, what is wanted is an output voltage, e_{10}, that varies in amplitude according to the instantaneous frequency of the incoming signal. Amplitude variations therefore must be removed ahead of the discriminator to keep their instantaneous values from adding to those values produced by the instantaneous frequency changes.

The input voltage, e_1, is applied across the input tuned circuit. The current, i_1, lags e_1 by 90°. The mutually induced voltage, e_2, lags i_1 by 90°. Thus, e_2 is 180° out of phase with e_1, as shown in part ① of figure 8–23, B.

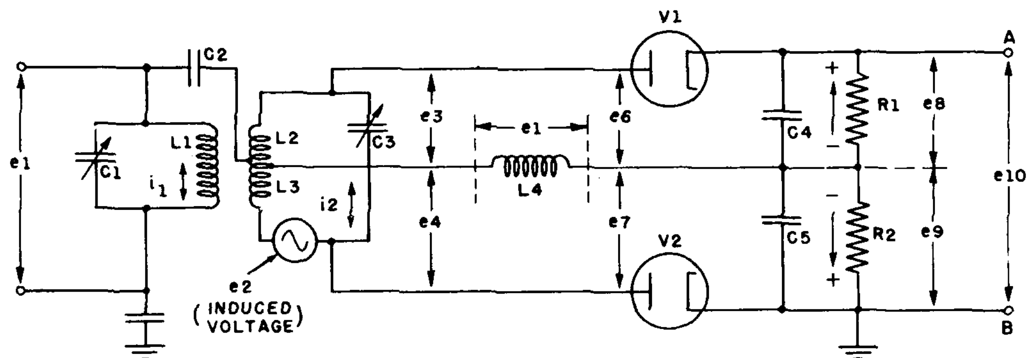
Inductor *L*4 is shunted across the input tuned circuit via *C*2 and *C*5, which have negligible reactance at the resonant frequency. Thus e_1 is also applied across *L*4.

Assume first that the incoming signal is at the resting frequency. The induced current, i_2, is in phase with e_2, as shown in parts ② and ③ of figure 8–23, B. The voltages e_3 and e_4 are the iX_L drops across *L*2 and *L*3 respectively. From figure 8–23, A, and part ③ of 8–23, B, it may be seen that e_6, the voltage applied to *V*1, is the vectorial sum of e_1 and e_3; and e_7, the voltage applied to *V*2, is the vectorial sum of e_1 and e_4. The rectified output voltage of *V*1 is e_8 and that of *V*2 is e_9. The output voltage, e_{10}, is the algebraic sum of e_8 and e_9. In part ③ of figure 8–23, B, e_6 and e_7 are equal because the incoming signal is at the resting frequency. Therefore e_8 is equal to e_9; and since they are in opposite directions, the output voltage is zero.

Below the resting frequency i_2 leads e_2 because X_C is greater than X_L. Voltages e_3 and e_4 are still in phase opposition, but each is 90° out of phase with i_2, as shown in figure 8–22, C. Therefore, e_7 is greater than e_6, and e_9 is greater than e_8. Point *A* becomes negative with respect to ground, thus producing an output signal voltage.

Above the resting frequency, i_2 lags e_2 because X_L is greater than X_C. Voltages e_3 and e_4 bear the same phase relation with each other and with i_2 as they did for each of the above conditions. However, from figure 8–23, D, e_6 is now greater than e_7. Therefore, e_8 is greater than e_9, and point *A* becomes positive with respect to ground, thus producing the other half of the audio signal-voltage waveform.

Ratio detector.—One form of the ratio detector is shown in figure 8–24, A. It differs from the discriminator in figure 8–23, A, in that the diodes are connected in series across the transformer secondary; whereas each diode plate in the



A

DISCRIMINATOR CIRCUIT

Figure 8-23.—Discriminator circuit and vector diagrams.

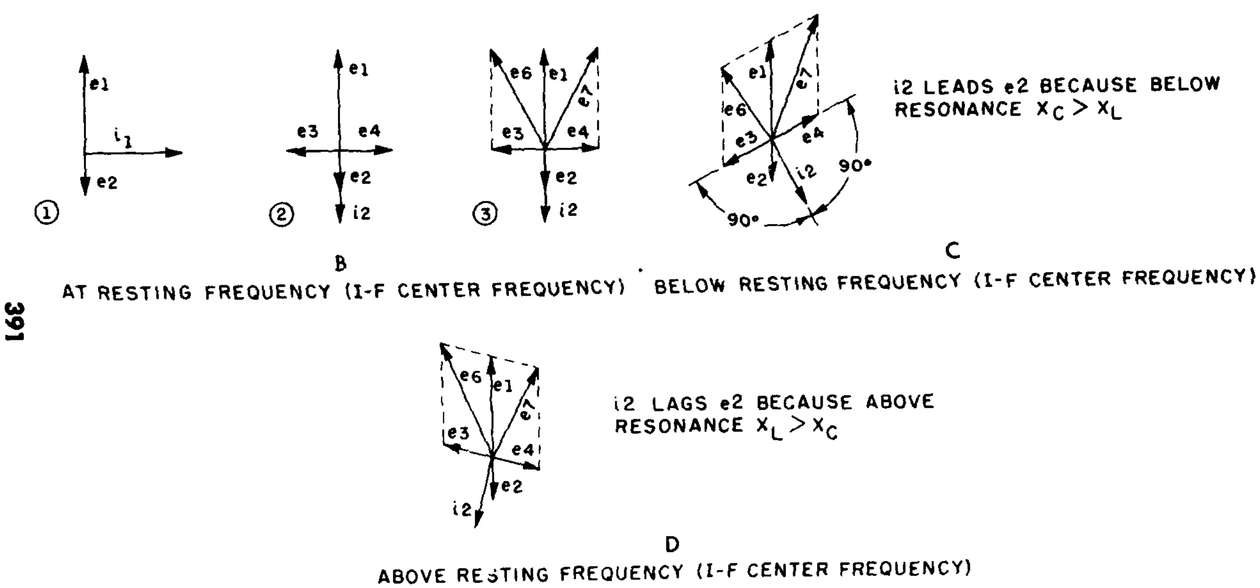


Figure 8-23.—Discriminator circuit and vector diagrams—Continued

discriminator is connected to opposite ends of the transformer. There are differences also in the method of obtaining the output voltage and in the amount of limiting preceding the detector. However, the vector analysis is essentially the same in both circuits.

The induced voltage, e_2, is 180° out of phase with e_1, as indicated by the vector diagrams in figure 8–24. At resonance (fig. 8–24, B), i_2 is in phase with e_2, as in all tuned circuits. Voltages e_3 and e_4, developed by the i_2X_{L2} and i_2X_{L3} voltage drops respectively, are 180° out of phase with each other and 90° out of phase with i_2. This out-of-phase relation holds true at resonance as well as below or above resonance.

From figure 8–24, B, e_5, the a-c voltage applied across $V1$, is the vector sum of e_1 (coupled through $C3$ between the center tap and ground) and e_3. Also, e_6, the voltage applied across $V2$, is the vector sum of e_1 and e_4. At resonance, e_5 and e_6 are equal, and there is only one path for the current to take—that is, from D through $V2$, $L3$, $L2$, $V1$, C, A, and back to D. The voltage across $R1$ is equal to the voltage across $C6$, and the voltage across $R2$ is equal to the voltage across $C7$. Therefore, no output voltage (e_7) is developed because no current flows through $R3$.

Capacitor $C8$, having a large capacity, charges to the potential existing between points C and D. Thus, e_{10} is always equal to e_5 plus e_6, if the small voltage drop in $V2$ is neglected. Because the matched capacitors, $C6$ and $C7$, are connected across $C8$, e_{10} is equally divided between them If there is a sudden undesirable increase in amplitude (frequency remaining constant), the potential between C and D will increase. However, the bridge circuit will still be balanced, and no appreciable current will flow through $R3$. Therefore, the output will not be affected.

Below resonance (fig. 8–24, C), i_2 leads e_2 because X_C is greater than X_L. Therefore, e_6, the vector sum of e_1 and e_4, is greater than e_5, the vector sum of e_1 and e_3. The current will now flow in two different circuits. It will flow from C through A, D, $V2$, $L3$, $L2$, $V1$, and back to C. It will also flow from C through A, B, $L4$, $L2$, $V1$, and back to C. The

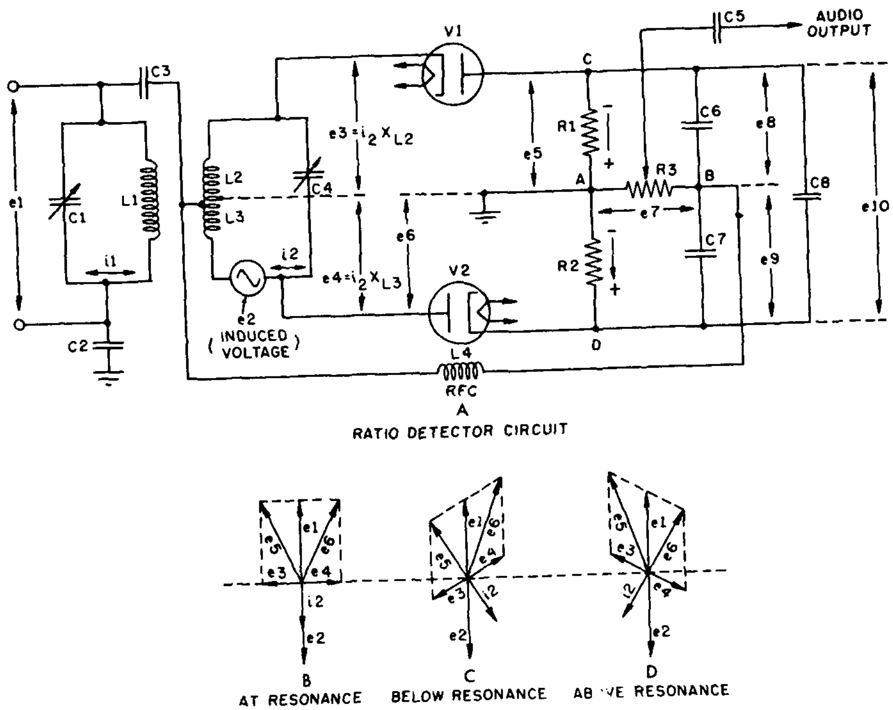


Figure 8-24.—Ratio detector and vector diagrams.

magnitude of the output voltage, e_7, developed by the flow of current through $R3$, depends on the RATIO between e_6 and e_5. When e_6 is larger more current will flow from A to B, and B will be more positive with respect to ground.

If the signal amplitude is suddenly increased, e_5 will be increased by a certain amount and e_6 will be increased by the same amount. Also, e_8 and e_9 will be increased in the same proportion. The bridge will therefore remain essentially in the same state of unbalance as it was before the sudden increase in amplitude was impressed between C and D. Therefore, an increase in amplitude of the input signal will not affect the amplitude of the output signal as much as a change in frequency of the input signal.

Above resonance (fig. 8–24, D), i_2 lags e_2 because X_L is greater than X_C. Therefore, e_5 is greater than e_6. The current will divide between two different circuits. It will flow from the center tap on the secondary through $L2$, $V1$, C, A, D, $V2$, and back to the center tap. It will also flow from the center tap on the secondary through $L4$, B, A, D, $V2$, and back to the center tap. The magnitude of the output voltage, e_7, developed by the flow of current through $R3$ from B to A depends on the ratio between e_5 and e_6. When e_5 is larger more current will flow from B to A, and B will be more negative with respect to ground.

A sudden undesirable increase in amplitude will increase the potential between C and D. Both e_5 and e_6 will be increased proportionately, and e_8 and e_9 will be increased in the same proportion. The bridge will therefore remain in essentially the same state of unbalance as before the sudden increase in voltage, and the output signal will not be affected as much as it would by a change in frequency.

Although the ratio detector requires fewer i-f amplifier stages and minimum limiting, it presents alignment difficulties, and the signal may be distorted at high input voltages if some form of limiting is not applied.

QUIZ

1. In what frequency components of the a-m wave is the intelligence contained?
2. What percentage of modulation corresponds to the condition of maximum permissible power in the side bands?
3. Why is the band of frequencies that may be transmitted on a-m restricted?
4. In the circuit shown in figure 8–7, how is the modulation signal fed to the *V*1 tank?
5. If the plate supply voltage is 1,000 volts in figure 8–7, what is the approximate peak voltage developed across the tank capacitor when the modulation is 100 percent?
6. For 100-percent modulation, what relation exists between the a-f and r-f input power?
7. In figure 8-7 what is the approximate maximum value of the plate voltage when the modulation is 100 percent?
8. When 100-percent modulation occurs, what is the percentage increase in antenna current over the unmodulated value?
9. What is the result of having the negative peak audio-modulation voltage exceed the B-supply voltage in figure 8–7?
10. What is the relative magnitude of the peak power output of the transmitter r-f amplifier during 100-percent modulation compared with that of the unmodulated peak power output?
11. What class of r-f amplifier is usually used with high-level plate modulation for maximum efficiency and ease of adjustment?
12. What are the advantages of grid modulation?
13. What are the disadvantages of grid modulation?
14. Why may a tone-modulated carrier be modulated approximately 100 percent?
15. Why are m-c-w signals less apt to be lost as a result of transmitter frequency drift than are c-w signals?
16. In f-m, upon what does the deviation frequency of the carrier depend?
17. How does modulation affect the energy distribution of the f-m wave?
18. Express the approximate bandwidth of an f-m signal in terms of the modulation frequency and the carrier frequency deviation.
19. Why may f-m transmitter tubes be operated at maximum efficiency at all times?
20. Why is frequency multiplication used in f-m systems?

21. Essentially, how does the reactance tube connected in parallel with the oscillator tank, function?
22. What is the purpose of the preemphasis circuit in a phase-modulated transmitter?
23. Define a nonlinear device.
24. What additional frequencies are introduced at a transmitter when the r-f carrier and the modulation frequency are impressed on a nonlinear device?
25. In what two ways are modulation and demodulation similiar?
26. Why is a square-law detector so called?
27. What is the approximate dividing line (in magnitude of voltage imput) between a weak-signal detector and a power detector?
28. Weak-signal detectors are always of what type?
29. In a diode detector, what keeps the plate current from decreasing to zero when the plate voltage approaches zero?
30. Why is a diode detector classified as a power detector?
31. Why must the capacitor across diode load resistor R in figure 8–18 be large enough so that the RC time constant is long compared with the time of one r-f cycle?
32. In figure 8–18, what places a limit on the highest modulation (audio) frequency that can be detected?
33. Name four disadvantages of the diode detector?
34. The grid-leak detector functions like what two electron-tube circuits?
35. Why does a grid-leak detector load the stage feeding it?
36. Why does the plate detector not load the preceding stage?
37. What is a disadvantage of slope detection of f-m signals?
38. Why must the discriminator be preceded by one or more limiter stages?
39. In the ratio detector of figure 8–24, upon what ratio of voltages does the magnitude of the output voltage across $R3$ depend?

CHAPTER 9

TRANSMITTERS

INTRODUCTION

A transmitter is a device for converting intelligence, such as voice or code, into electrical impulses for transmission either on closed lines, or through space from a radiating antenna. Transmitters take many forms, have varying degrees of complexity, develop various levels of power, and employ numerous methods of sending the desired information or energy component from one point to another.

A telephone handset has both a transmitter and a receiver section. The transmitter section converts the human voice into electrical impulses that may be amplified and conveyed along the closed telephone line to the receiving station.

Radar transmitters develop bursts of energy of the required frequency and duration and radiate this energy in the direction in which the antenna is pointing. The echo may be utilized to give such information as range and bearing. The tremendous energy radiated by some radar sets is made possible because of the relatively long resting time between pulses of energy. Radar will be treated briefly in chapter 14.

Loran transmitters are especially constructed for use in navigation. The principle of loran is based on the difference in time required for pulsed radio signals to arrive at a point from a pair of synchronized transmitters. These transmitters operate on low frequencies, are accurately synchronized,

and develop considerable power because they are transmitting only a small fraction of the time. Loran transmitters, as well as sonar and other transmitters, are treated in the rating texts.

In this chapter the discussion is confined to radio transmitters; however, many of the principles involved in this discussion apply in general to other basic electronic transmitters.

The function of a radio transmitter is to supply power to an antenna at a definite radio frequency and to convey intelligence by means of the radiated signal. Radio transmitters radiate waves of two general types.

One type of radiation is the CONTINUOUS WAVE (c-w), or UNMODULATED WAVE, which has a waveform like that of the r-f current in the tuned tank circuit of a power output stage. In this type of wave the peaks of all the waves are equal, and they are evenly spaced along the time axis. The waveform is sinusoidal.

The other type of radio wave is the MODULATED WAVE. The amplitude may be modulated by means of a signal of constant frequency, as in MODULATED-CONTINUOUS-WAVE (m-c-w) telegraphy. Likewise, the amplitude may be modulated by means of speech, music, and so forth; and in this case it is called AMPLITUDE MODULATION (a-m). If the frequency of the wave is varied with time it is called FREQUENCY MODULATION (f-m). Although there are other types of modulation—for example, pulse-time modulation—only c-w, a-m, and f-m will be treated in this chapter.

A given transmitter operated on c-w has a greater range than the same transmitter (for the same power output) operated on m-c-w or voice modulation. This condition results from the fact that all the intelligence is contained in the side bands (treated in chapter 8), and the fewer the number of side-band frequencies the greater will be the signal strength in the remaining side-band frequencies. In c-w operation the side bands do not extend very far on each side of the carrier, and all of the energy is therefore contained in a narrow band and not wasted in nonessential bands.

On m-c-w the side bands are necessarily wider, and more energy is needed to supply the side bands; that is, each side band contains proportionately less energy. In order to get the same signal level (at the required bandwidth) to a receiver, the transmitter must increase its output power over what it would be if c-w were used.

When voice modulation is used, the necessary side bands are increased over those needed for m-c-w. Each side band requires a certain amount of energy, and therefore, in order to keep the energy level of all the essential side bands up to the required level at the receiver, the transmitter must deliver more energy than for m-c-w.

Navy transmitters operate on very-low-frequency (v-l-f), low-frequency (l-f), medium-frequency (m-f), and high-frequency (h-f) bands as well as on very-high-frequency (v-h-f) and ultrahigh-frequency (u-h-f) bands.

The VERY-LOW-FREQUENCY BAND, from 10 to 30 kilocycles, is not covered by shipboard transmitters. The antennas needed for such low frequencies are too long to be erected aboard a ship. However, there are some v-l-f stations on shore. One of the frequencies of the Primary Fleet Broadcast, NSS, is in the very-low-frequency band. Powerful v-l-f stations with their huge antennas are capable of transmitting signals through magnetic storms that blank out the higher radio-frequency channels. One such station (the most powerful to date) has recently been erected near Seattle, Washington.

The LOW-FREQUENCY BAND, from 30 to 300 kc, is used mostly for long-range direction finding. This band provides, however, a means of reliable medium- and long-range communication. Useful amounts of radiation can be produced in this band with shipboard antennas. The frequencies in the low-frequency band do not depend on sky waves and provide stable communication with little variation from season to season.

The area that may be covered by the MEDIUM-FREQUENCY BAND, extending from 300 to 3,000 kc, depends on the ground wave. Sky-wave reception of medium-frequency waves is

also possible. (Propagation is treated in chapter 11.) At the upper end of this band the ionosphere has a great effect on the sky waves. Relatively long distances can be covered by using this band if the correct frequency is used at the correct time. The international distress frequency, 500 kc, is in this band. Commercial broadcast stations as well as Navy stations operate in this band.

The HIGH-FREQUENCY BAND is also used by the Navy. This band includes the frequencies between 3 and 30 megacycles. The sky wave is increasingly important in this band, and long-range communication is possible. Propagation characteristics of waves in the h-f band change with the time of day and the season. The choice of frequency depends on the variables in the ionosphere. For long-range ship-to-ship and ship-to-shore communications either the h-f band or the upper part of the m-f band may be employed, depending mostly on the time of day. Transmitters such as the TBM, TBL, TBK, and TDE, cover part of the m-f and most of the h-f bands. The TBL and TDE also cover a portion of the l-f band.

The VERY-HIGH-FREQUENCY BAND extends from 30 to 300 megacycles. This band is not used extensively by the Navy for communications purposes. However, the 60- to 80-mc band is used by Navy type TBS and MBF transmitters, and the 100- to 150-mc band is used by Navy type TDQ and the AN/ARC series of transmitters. Both of these bands are used largely for emergency purposes. Portions of the v-h-f band, however, are used for airborne communications. The TDQ and TBS transmitters and the MBF transmitter-receiver operate in this band. Dependable communications at distances slightly greater than the horizon can be obtained with these equipments. The v-h-f band is used with early warning radar, IFF, television, and f-m broadcast stations.

The ULTRAHIGH-FREQUENCY BAND includes the frequencies between 300 and 3,000 megacycles. The low end of this band is used for communications and portions of the high end are used for radar. Navy type u-h-f transmitters include the TDZ, and TED, the MAR, and the AN/ARC 27,

all of which have frequency ranges from 200 to 400 mc (extending into the v-h-f band).

Thus it may be seen that Navy transmitters must be flexible as far as type of modulation (c-w, m-c-w, or voice) and frequency range (frequency multiplication is often used to extend the range) are concerned. The maximum in utility must be obtained from the equipment for a minimum in size and weight. Therefore, care in design and modifications where needed characterize these transmitters. A few of the more important features that must be incorporated in every Navy transmitter are excellent frequency stability, ruggedness, long life, flexibility of operation, remote-control operation, ease of tuning, and high efficiency. Individual transmitters often have several modifications, the main difference being in the power supplies or in minor mechanical or electrical changes. Navy equipment is somewhat different from corresponding commercial equipment. Some of the differences may seem of little significance, but over a period of years the equipment designed specifically for Navy use has proved superior.

CONTINUOUS-WAVE TRANSMITTERS

Introduction

The continuous wave is used principally for radio telegraphy—that is, for the transmission of short or long pulses of r-f energy to form the dots and dashes of the Morse code characters. C-w transmission was the first type of radio communication used, and it is still used extensively for long-range communications. Some of the advantages of c-w transmission are narrow bandwidth; high degree of intelligibility, even under severe noise conditions; long range; and security, because the message may be encrypted.

A modern application of c-w transmission is the teletype, which replaces the operator for changing the plain language into code and code into plain language. To transmit a message with teletype the operator presses keys on a keyboard

similar to a standard typewriter. As each key is pressed, mechanical cams and linkages cause a sequence of mark and space signals to be sent out by the transmitter. At the receiving end, the received signal actuates selector magnets in a similar machine that causes the character transmitted to be printed on a paper and the carriage to be advanced one space. The code characters may be cut in a tape by the operator at relatively slow speeds and later transmitted at high speeds. Automatic transmission of messages has the advantages of speed and rythmic character transmission for easier manual copying by an operator. However, it has the great disadvantage of being very easily interfered with. When interference is too great, an operator must take over and transscribe on a typewriter.

The four essential components of a c-w transmitter are: (1) a generator of r-f oscillations, (2) a means of amplifying these oscillations, (3) a method of turning the r-f output on and off (keying) in accordance with the code to be transmitted, and (4) an antenna to radiate the keyed output of the transmitter. A block diagram of a master-oscillator power-amplifier transmitter together with the power supply is shown in figure 9–1.

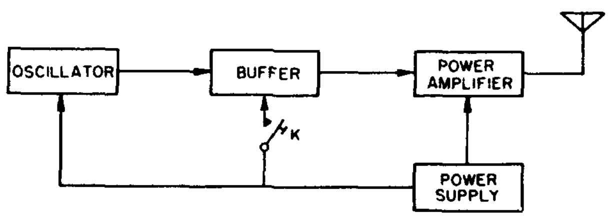


Figure 9–1.—Master-oscillator power-amplifier transmitter.

Oscillator

One of the most important sections of a transmitter is the one containing the oscillator. Here the frequency on which the transmitter operates (or a subharmonic of the transmitter frequency, if frequency doublers or multipliers are used) is

generated and maintained within the required limits. Two of the problems encountered in this section of the transmitter are (1) maintaining a stable frequency, and (2) switching from one frequency to another with a minimum of adjustments. Oscillators are treated in chapter 7. Only some of the special problems as applied to transmitters will be considered in this chapter.

One of the major problems encountered in the operation of transmitters is that of frequency drift. The frequency of a transmitter should be stable enough to avoid wandering into another band and causing interference; also, it should be stable enough to permit a receiver to stay on the transmitter frequency. However, the master oscillator in a transmitter tends to change frequency when it is being warmed up and when the load on the oscillator varies. The law requires that the carrier frequency be held very close to the specific frequency assigned by the Federal Communications Commission. For example, the frequency tolerance allowed an international broadcast station is 0.005 percent of the assigned frequency.

The frequency of a transmitter can be stabilized by the use of a crystal oscillator However, this arrangement would require a large number of crystals to cover the many frequency channels used by the Navy. A more flexible means of obtaining stability (especially at the lower frequencies) is to control the frequency of a transmitter with a variable master oscillator. An electron-coupled oscillator (ECO) is commonly used.

Most of the frequency drift in master oscillators is due to changes in the physical size of the components with variations in temperature; therefore, changes in the electrical characteristics of the oscillator circuit, as well as changes in the oscillator tube characteristics, are introduced. Placing the frequency-determining components of the oscillator in a temperature-controlled oven eliminates this drift. To ensure further stability the oscillator is loaded very lightly and isolated by a buffer stage.

The frequency of the master oscillator can be affected also

by vibration and sudden shocks. In some transmitters all of the oscillator elements are mounted in a single oscillator unit. The oscillator unit is then suspended on springs and snubbed by sponge rubber cushions to keep the shock and vibrations reaching the oscillator unit to a minimum.

Frequency stability becomes even more important when a transmitter uses frequency multiplier stages because any drift in the oscillator frequency will be multiplied in these stages. For example, if the output frequency is eight times the oscillator frequency, any drift of the oscillator frequency will be multiplied by eight.

Not all Navy transmitters use all of the refinements, but every transmitter has some means of ensuring frequency stability.

The majority of master oscillator circuits in low- and medium-frequency Navy transmitters are electron-coupled oscillators because of their stability. The frequency of the oscillator is varied by either a variable capacitor or a variable inductor. Different frequency ranges are obtained by using a tapped oscillator coil or by switching in various values of capacitance. Sometimes both methods are used together. Usually the frequency of the oscillator is doubled in the plate circuit. With this arrangement, any energy fed back to the grid circuit is twice the frequency of the energy in the grid circuit and does not affect the stability of the oscillator.

Most of the transmitters operating in the u-h-f band are crystal controlled. However, crystals having a fundamental frequency in the u-h-f band are not practical because of the physical restrictions, such as the difficulty in grinding the crystals and their extreme fragility. They are about the size of a thin dime. Therefore, the transmitter usually employs a low-frequency crystal oscillator followed by a number of frequency multipliers, the number used depending on the desired output frequency. Of course, the multiplier stages must be accurately tuned to the correct harmonic frequency. In this arrangement the crystal is larger and more substantial than it would be if it were operated at the higher frequencies.

Buffer Amplifier

As mentioned previously, a buffer amplifier is placed between the oscillator and the power amplifier to isolate the oscillator from the load and thus improve the frequency stability of the transmitter. If the frequency of the plate tank circuit of the buffer amplifier is the same as that of the oscillator driving it, the stage is a conventional type of amplifier, usually class C.

If the plate tank circuit of the buffer amplifier is tuned to the second harmonic (in order to increase the frequency of the radiated signal) of the driving signal applied to the grid, the stage becomes a frequency doubler and the output voltage has a frequency equal to twice that of the input. Likewise, the buffer amplifier may become a tripler or a quadrupler.

A frequency-doubler stage is shown in figure 9–2, A. The plate tank is tuned to twice the frequency of the grid tank. If $L1$ is equal to 10 μh and $C1$ is equal to 25.3 $\mu\mu$f, the resonant frequency of the grid tank is

$$f_1=\frac{159}{\sqrt{LC}}=\frac{159}{\sqrt{10\times 25.3}}=10 \text{ mc.}$$

If the plate tank coil has an inductance of 10 μh and the resonant frequency of the plate tank is 20 mc, the plate tank capacitor, $C2$, will have a value of

$$\frac{C1}{4}=\frac{25.3}{4}=6.3\ \mu\mu f.$$

The capacitance of $C2$ may be verified by substituting the values of f and L in the following formulas:

$$LC=\left(\frac{159}{f}\right)^2;$$

thus,

$$10C=\left(\frac{159}{20}\right)^2,$$

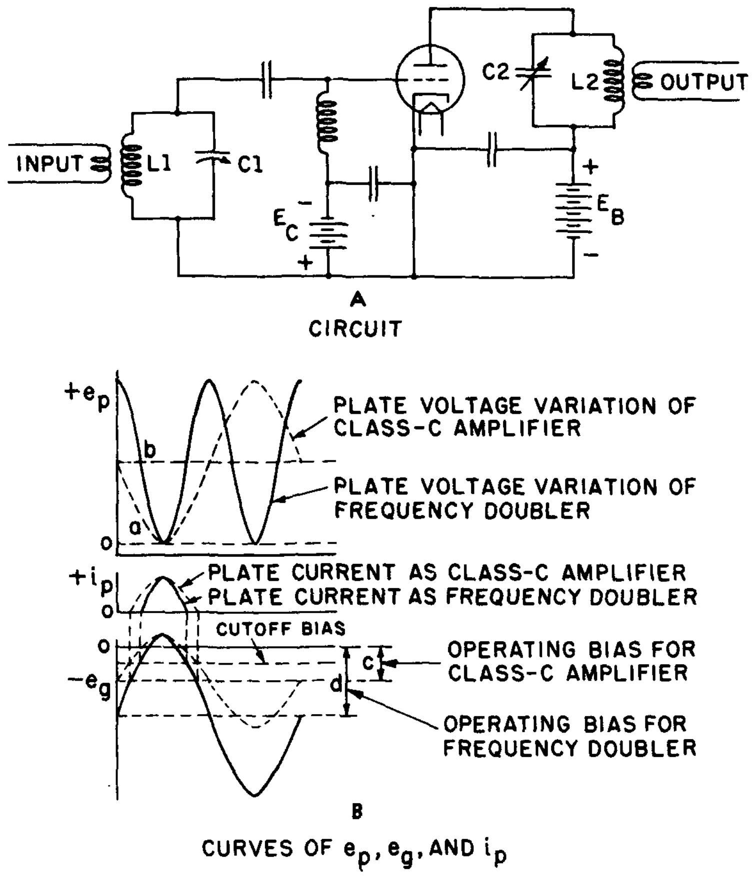


Figure 9–2.—Frequency doubler

from which

$$C=\frac{159^2}{20^2\times 10}=6.35\ \mu\mu f.$$

The curves of plate voltage, grid voltage, and plate current are shown in figure 9–2, B. The dotted curves indicate operation as a class-C amplifier without frequency multiplication, and the solid curves indicate operation as a frequency doubler. Unless the operating bias is increased,

the triode plate will overheat when the plate tank is tuned to the second harmonic of the grid-tank circuit. Plate voltage is higher (*a* to *b*) during the interval the grid voltage is above cutoff, and the duration of plate current flow is reduced by increasing the operating bias from *c* to *d*.

Although the angle of plate-current flow is reduced, the efficiency is maintained. The output of the frequency multiplier varies inversely with the extent of frequency multiplication. If the plate tank is tuned to the second harmonic of the grid tank, the duration of flow of plate current is from 90° to 100° and the power output is about 65 percent of the output of a class-C amplifier. If the plate tank is tuned to the third harmonic of the grid tank, the angle of flow of plate current is from 80° to 120° and the output is reduced to 40 percent of that of a class-C amplifier. If the plate tank is tuned to the fourth harmonic of the grid input signal, the angle of plate current flow is reduced to between 70° and 90°, and the output is 30 percent of that of a class-C amplifier. If the frequency is multiplied by 5, the angle of plate current flow is from 60° to 72° and the output power is 25 percent of that of a class-C amplifier.

In every case it is necessary to increase the operating bias and the grid driving signal as the frequency multiplication increases in order not to overheat the triode plate. The flywheel effect in the plate tank supplies the missing cycles of grid drive and the output is approximately an undamped wave having sine waveform.

Three important conditions must prevail in order to obtain frequency multiplication—(1) high grid-driving voltage, (2) high grid bias, and (3) plate tank tuned to the desired harmonic. If the second harmonic is selected, the stage is called a FREQUENCY DOUBLER; if the third is used, the circuit is called a FREQUENCY TRIPLER, and so forth.

Certain amplifier circuits are suited to the generation of even harmonics and others to the generation of odd harmonics. Push-pull amplifiers produce only odd harmonic frequency multiplication—third, fifth, seventh, and so forth. If the grids of two triodes are connected in push-

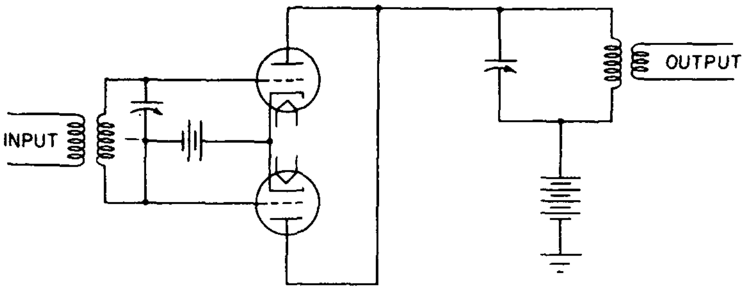


Figure 9–3.—Even-order harmonic frequency multiplier.

pull and the plates in parallel (fig. 9–3), even-order harmonics can be produced.

The grid signals are 180° out of phase. When one grid voltage is positive maximum, the other is negative maximum, and the second alternation of the cycle reverses the respective potentials. Thus pulsating plate current flows first in one tube and then in the other. Because the plates are connected in parallel, the output pulses are in the same direction and the plate tank circuit receives two pulses for each input cycle at the grids. This type of doubler is capable of greater output and higher plate efficiency than the single-tube type.

Power Amplifier

Wherefore the buffer stage isolates the oscillator from the varying load caused by the keying, the power amplifier increases the magnitude of the r-f current and voltage by the resonant action of the plate tank circuit. The power amplifier shown in figure 9–4 is a class-C amplifier. The triode amplification factor, μ, is 20 and the plate supply voltage is 1,000 volts. The cutoff bias, e_{co}, is therefore

$$e_{co}=\frac{-e_p}{\mu}=\frac{-1{,}000}{20}=-50 \text{ volts.}$$

The operating bias, e_o, is three times the cutoff, and therefore

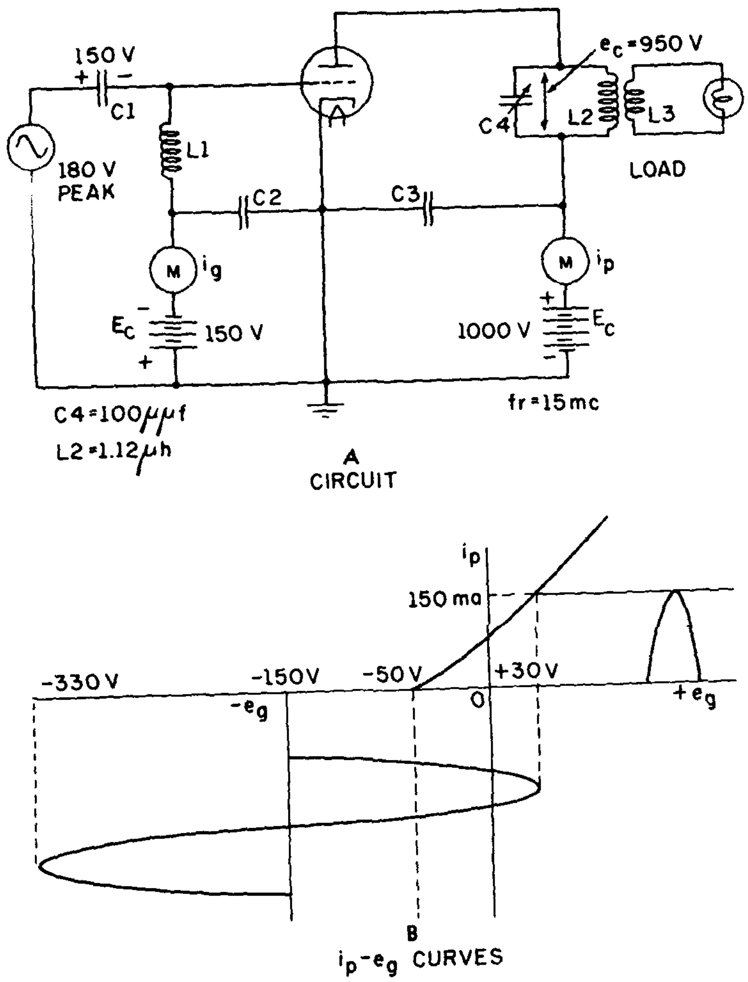


Figure 9–4.—Power amplifier.

$$e_o = 3(-50) = -150 \text{ volts.}$$

The maximum value of the r-f input signal is 180 volts. Thus, when the grid end of the r-f input is positive, the peak positive grid-to-cathode voltage is $180-150=+30$ volts. When the grid end of the r-f input is negative, the peak negative grid-to-cathode voltage is $(-180)+(-150)=-330$ volts.

When the grid voltage is above the cutoff value, plate current flows, and at the instant the grid voltage is +30 volts, the plate current is 150 ma (fig. 9–4, B). The tuning capacitor, $C4$, charges up to nearly the full value of the B-supply voltage, or 950 volts. During this charging process, the lower capacitor plate is positive and the upper plate is negative. Thus, the instantaneous triode plate-to-ground voltage, when the capacitor voltage is 950 volts, is $1{,}000-950=50$ volts. This value is called e_{min} and represents the lowest value of plate-to-cathode voltage in the entire cycle.

The relations between plate voltage, plate current, grid excitation voltage, and resonant plate tank circuit voltage and current are shown in figure 9–5. The flywheel effect in the plate tank circuit causes the capacitor to periodically reverse its polarity and continue the a-c cycle within the tank when the grid voltage is below cutoff and no energy is being supplied to the tank circuit from the power supply. The plate voltage swings from 1,000 volts to a minimum of 50 volts and then to a maximum of 1,950 volts before it completes the cycle.

Thus, the plate tank circuit converts pulses of unidirectional current in the triode plate circuit into sine-wave variations of current in the resonant tank. These surging currents give the tank circuit the so-called FLYWHEEL EFFECT, in which the tuned circuit makes up the portion of the sine wave missing in the plate current pulses, and supplies a voltage of sine waveform to the load. The tube acts as a valve merely to supply the necessary power at just the right time.

Plate current flows for about one-third of each cycle.

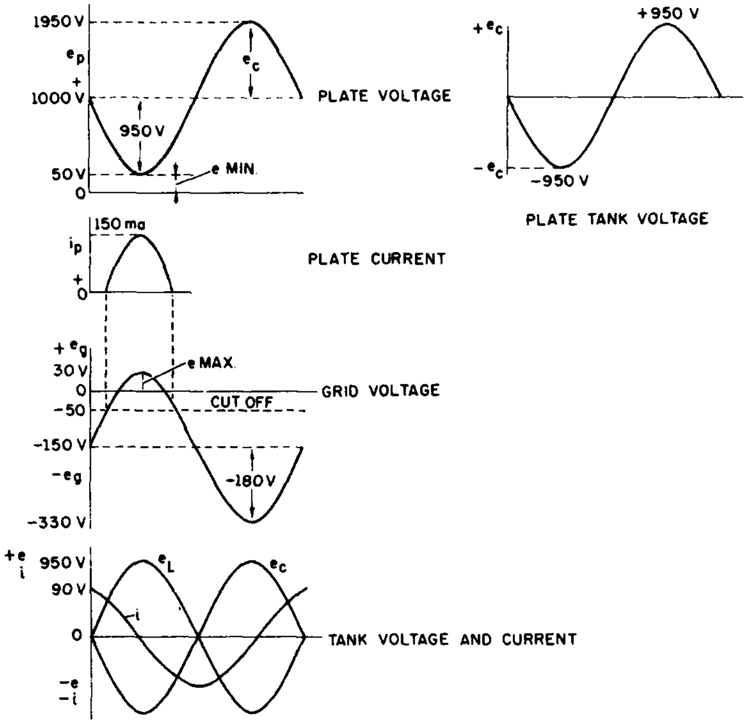


Figure 9-5.—Power amplifier current and voltage relation.

Energy is supplied to the tank circuit during the time plate current flows. Plate voltage, e_p, is below the power-supply value during this portion of the cycle because the tank capacitor charges up with a polarity that opposes the polarity of the power supply, and $e_p = E_b - e_c$, where e_c is the voltage across the tank capacitor. Thus, energy is supplied to the tank with minimum plate losses because it is supplied at a time when the plate voltage (hence plate losses) is at a minimum. The efficiency of the class-C amplifier may be as high as 70 percent.

Grid voltage is positive with respect to the cathode for

a short time in each cycle, and for optimum conditions the minimum value of plate voltage, e_{min}, should be equal to the maximum positive value of grid driving voltage, e_{max}. In other words, $e_{max}=e_{min}$. The maximum positive grid voltage, e_{max}, should never be allowed to exceed the minimum plate voltage, e_{min}. Otherwise, plate current would decrease and grid current would become excessive resulting in a reduction in output power and excessive grid losses. Thus in figure 9–5, e_{max} is made 20 volts less than e_{min} in order to ensure that e_{max} will never exceed e_{min}.

The plate tank circuit in figure 9–4 may have an artificial load applied to it for the purpose of tuning the amplifier prior to coupling the antenna to it. This load may take the form of an incandescent lamp of approximately the same power rating as the amplifier supplying it. The load is coupled to the tank by means of the link coil, *L*3.

The plate supply voltage is reduced and tuning capacitor *C*4 is adjusted for resonance, as indicated by the dip in the plate milliammeter (line current is a minimum at resonance). The sharp decrease in plate current is accompanied by a corresponding increase in tank current. As resonance is approached, grid current increases as plate current decreases. The load on the tank circuit may be increased by moving link coupling inductor *L*3 closer to tank coil *L*2 and increasing the plate supply voltage to the normal value.

The increased load on the amplifier increases the current through the lamp and decreases the current in the plate tank circuit. The decrease in current in the resonant tank is accompanied by a decrease in voltage, e_c, across the tank. Thus in figure 9–5, e_{min} becomes larger ($e_{min}=E_b-e_c$) and plate current increases with the load. The space current increases because plate voltage is increased during that portion of the cycle when the triode is conducting. For a given filament emission, plate current increases and grid current decreases as the load on the tank increases.

The transmitter output may be coupled to the antenna by tuning the antenna to resonance and coupling it to the final amplifier tank by means of a link coil similar to *L*3.

The artificial load is removed by moving $L3$ away from $L2$, as the antenna load is applied to the tank.

Bias Methods

Most of the bias methods used in receivers can also be employed in transmitters. However, because of the power output requirements of transmitters, class-B or class-C r-f amplifiers are used most often and these call for grid-leak bias. Grid-leak bias depends on grid current flow for a portion of the input cycle. This type of bias is not generally used in receivers because most r-f amplifiers in receivers operate class A with no grid current.

A grid-leak bias circuit is shown in figure 9-6. The triode is assumed to operate as a class-C amplifier with a peak driving voltage of 180 volts, a cutoff bias of -50 volts, and an operating bias of -150 volts. The polarities and magnitudes of the voltages for the condition of maximum positive grid-to-cathode voltage are shown in figure 9-6, A. The polarities and magnitudes of the voltages for the condition of maximum negative grid-to-cathode voltage are shown in figure 9-6, B. The wave forms of the grid-driving voltage and plate current are shown in figure 9-6, C.

Capacitor $C1$ blocks grid current from the signal source. Capacitor $C2$ is an r-f bypass capacitor that holds the lower end of the r-f choke at r-f ground potential. Grid-leak bias voltage is developed across grid resistor R.

In this example, R has a resistance of 15 k-ohms and the grid current is 10 milliamperes. The voltage across R is 10×15, or 150 volts. The voltage across R without $C2$ is a series of half-wave pulses. The fact that $C2$ is shunted across R smoothes these pulses into a steady d-c bias voltage. The electron flow through R makes the grid end negative and the cathode end positive.

The a-c driving voltage is developed across the r-f choke which presents high inductive reactance to the r-f input and low effective resistance to the d-c grid current. The capacitive reactances of capacitors $C1$ and $C2$ are low so that the a-c voltage across them is negligible.

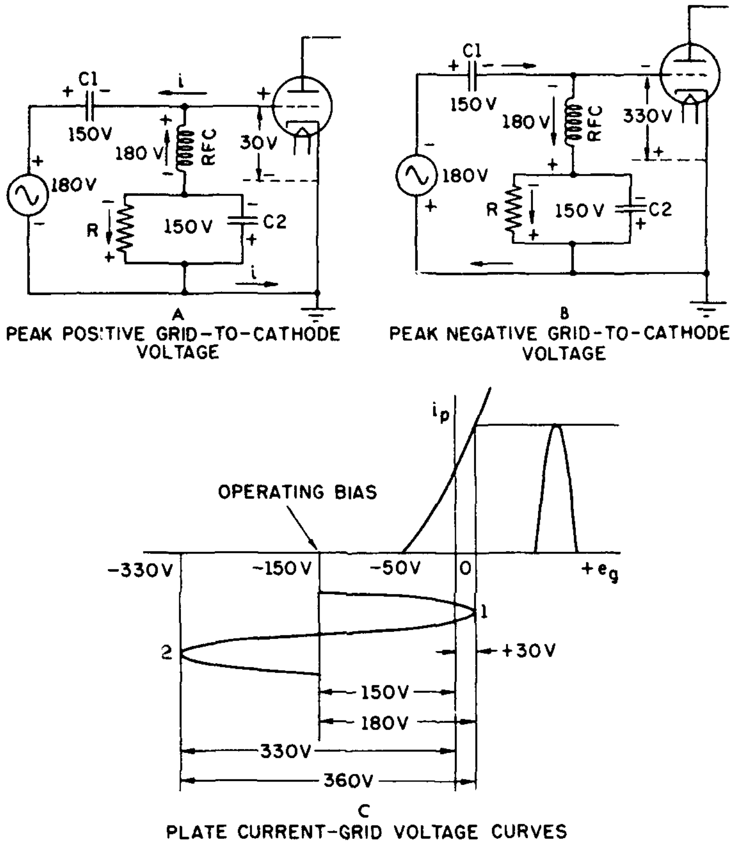


Figure 9-6.—Analysis of grid-leak bias.

In figure 9-6, A, the grid conducts and the positive peak grid-to-cathode voltage is 180−150, or +30 volts. During the time grid current flows, *C*1 charges up to 180−30, or 150 volts, and the low impedance of the conducting cathode-to-grid circuit bypasses grid current around the r-f choke and resistor *R*.

Also during the time the grid is conducting, *C*2 discharges through *R*, thus maintaining the operating bias of −150

volts. $C2$ is sufficiently large that its voltage does not change appreciably during discharge.

On that portion of the input cycle when the grid is negative with respect to the cathode, no grid current flows. At the instant pictured in figure 9–6, B, the grid is maximum negative with respect to the cathode. The path for the a-c input voltage is to the right through $C1$ and down through the r-f choke and grid resistor R. Capacitor $C2$ charges up as $C1$ discharges. $C1$ is sufficiently large that its d-c potential does not fall appreciably during discharge.

Grid-leak bias has the desirable characteristic of adjusting its value automatically when the amplitude of the grid driving voltage varies in magnitude. For example, an increase in driving voltage increases the operating bias, which checks the increase in grid current; or, if the grid driving voltage decreases, the decrease in grid current is checked by a shift of the operating bias in a positive direction. The correction is automatic in either case because it is the flow of grid current through the grid resistor that produces the operating bias. Thus the grid current is maintained at the proper value automatically over an appreciable range of input voltage.

In order to maintain grid-leak bias, grid current must flow a part of each cycle. Removing the driving voltage or lowering its amplitude below the value that drives the grid positive causes a loss in grid current and operating bias. Plate current then becomes dangerously high and the tube may be damaged.

Separate bias may be employed to prevent the condition of excessive plate current when grid bias is removed. Figure 9–7 shows a circuit that uses protective bias in series with grid-leak bias. In this example, the grid bias is developed as a result of the flow of 10 ma of grid current through a 10 k-ohm resistor. The voltage across R is 10×10, or 100 volts. The protective bias is provided by a 50-volt battery. Cutoff bias for the triode is assumed to be -50 volts. The total operating bias is $-100-50$, or -150 volts.

Grid current flows down through R and into the negative

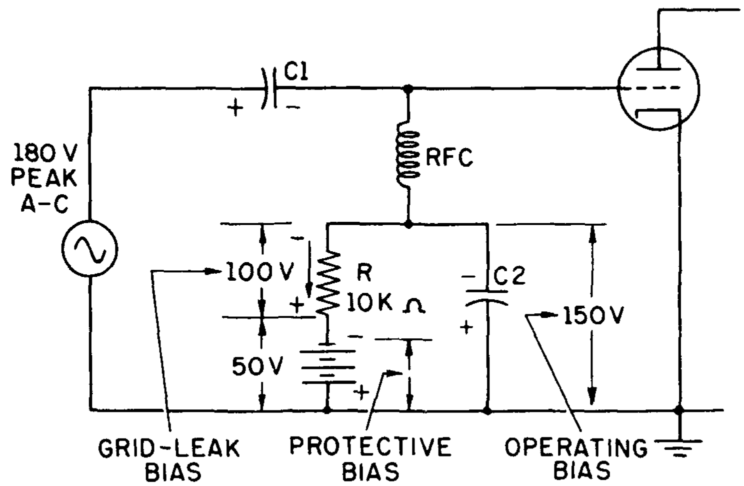


Figure 9–7.—Combination grid-leak and battery bias.

terminal of the battery. The direction is such as to charge the battery and raise its terminal voltage when grid current flows. The energy to charge the battery comes from the a-c driving voltage source. Thus, the voltage regulation of a separate bias supply, such as the bias battery, is negative (the greater the grid current, the higher is the bias supply voltage).

Other forms of protective bias, including cathode bias, and separate bias supplies are discussed in chapter 3.

Transmitter Electron Tubes

Electron tubes used in transmitters differ very little from those used in receivers except in size. Because most transmitter tubes are power tubes designed to amplify high voltage and heavy current, they must be of much larger and heavier construction. The plate dissipation of a tube is the difference between the plate power input and the power output. If this dissipation is greater than normal, the plate will become very hot, sometimes glowing with a red color from this heat. If the heat becomes intense, gases may

develop within the tube, making it unsatisfactory for further use. A transmitter should not be operated for any appreciable time if the plates become red unless the service manual for the set states that this condition is normal for

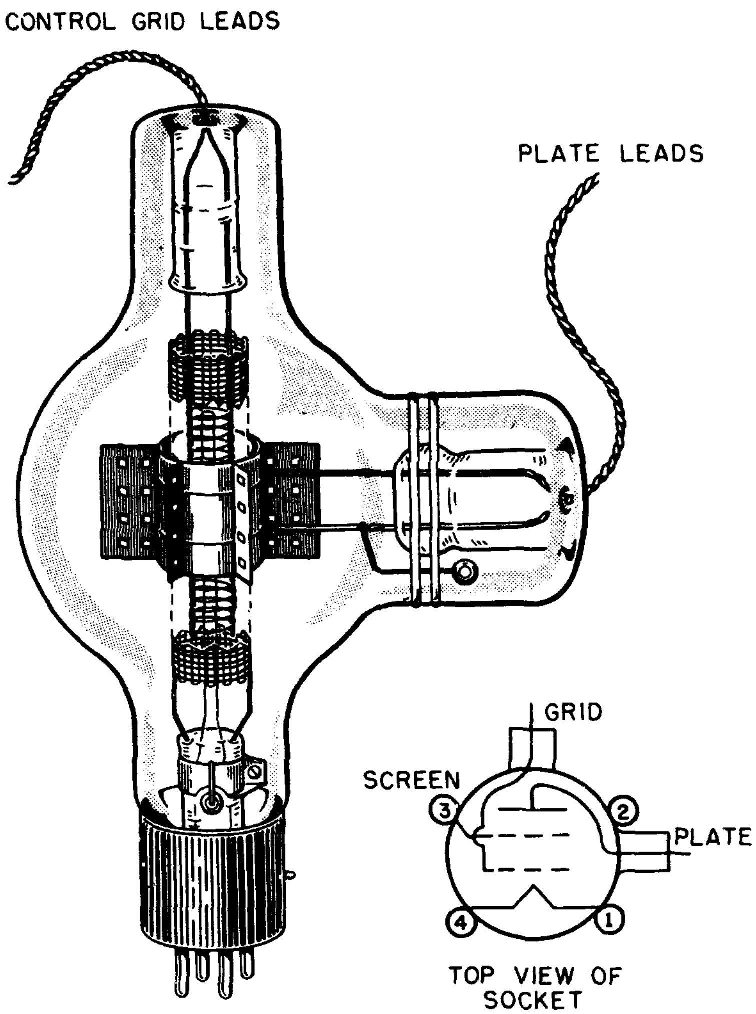


Figure 9–8.—Medium-power transmitting tetrode.

operation. Loss of bias, insufficient grid excitation, or improper tuning may cause overheating of a transmitter tube.

A common screen-grid tube similar to the type 860 used as a class-C r-f power amplifier in Navy transmitters is shown in figure 9–8. It requires no neutralization and is used in medium-power transmitters.

The thoriated tungsten filament has a rating of 10 volts and 3.25 amperes. The two filament leads and the screen-grid lead are brought out through the base pin connections. The screen-grid voltage is 300 volts. The control grid is connected to stranded leads that are brought out through a separate seal in the upper arm of the bulb. Control-grid voltage for class-C operation is −150 volts and grid current is 15 milliamperes. The driving power required is about 7 watts. The plate is connected to stranded leads that are brought out through a separate seal in the side arm of the bulb. Plate voltage for class-C operation is 3,000 volts.

This tube is mounted in a vertical position. The plate shows a dull red color when it is operated at the maximum plate dissipation rating of 100 watts. Normal plate current is 85 ma when the power output is 165 watts. Plate efficiency of this amplifier is about 65 percent when the plate dissipation is 90 watts. The amplification factor of this tetrode is 200, its transconductance is 1,100 micromhos, and its a-c plate resistance is approximately 182,000 ohms.

Because plate current flows for only a portion of each cycle, tubes are better able to dissipate the heat developed and thus have a longer life if they are operated class C. Additional details on the construction and operation of electron tubes are given in chapter 2.

Neutralization

A transmitter r-f amplifier having a plate tank and grid tank circuit both tuned to the same frequency resembles a tuned-plate tuned-grid oscillator. Unless some precaution is taken to prevent it, the amplifier may break into oscillation, causing a very unstable operating condition. If a

voltage is fed back from plate to grid in phase with the grid signal, oscillation will occur. If the voltage fed back is 180° out of phase with the grid signal, the action is degenerative and oscillations will be stopped.

Neutralization is a process of balancing the voltage fed back by the interelectrode capacitance of the tube with an equal voltage of opposite polarity. Dividing the plate circuit so that the neutralization voltage is developed across a part of it is called PLATE NEUTRALIZATION. If the voltage of neutralization is developed in the grid circuit the arrangement is called GRID NEUTRALIZATION.

In figure 9–9, the plate tank coil has a center-tap connec-

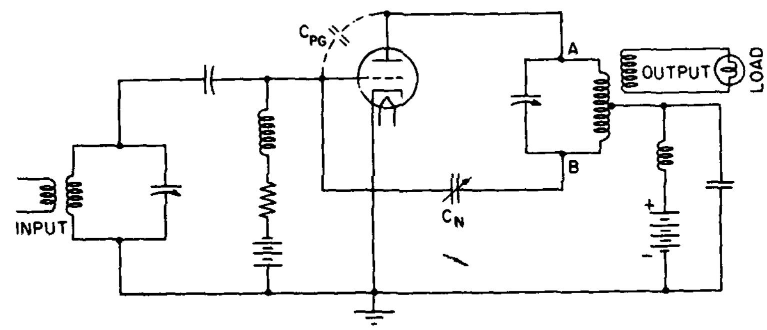


Figure 9–9.—Plate neutralization.

tion. The voltage between point A and ground is 180° out of phase with the voltage from point B to ground. Feedback through the plate-to-grid capacitance of the triode produces a voltage across the grid input circuit that is in phase with the grid excitation voltage and therefore tends to cause the amplifier to break into oscillations.

The neutralizing capacitor, C_N, couples a portion of the voltage between point B and ground to the grid input circuit. This action is degenerative and tends to block oscillations. A simplified equivalent circuit is shown in figure 9–10.

By the adjustment of C_N, the voltage fed back to the grid through C_N is made equal to the voltage fed back through the tube.

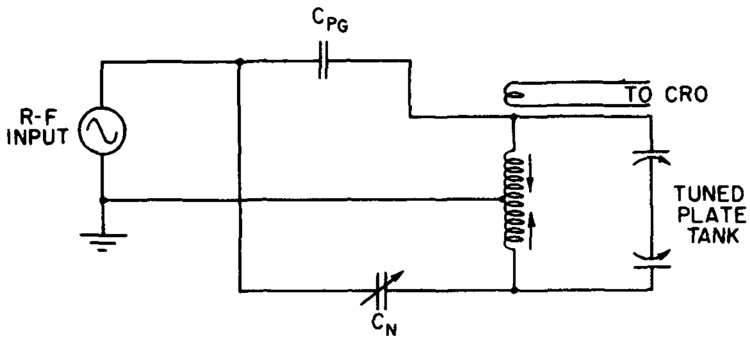


Figure 9–10.—Equivalent circuit for plate neutralization.

One method of determining the correct adjustment for C_N is to apply the input r-f voltage with normal filament but no plate voltage. A pick-up coil near the plate tank is fed to the vertical input of a cathode-ray oscilloscope. C_N is adjusted so that no r-f voltage appears on the scope when the plate tank is tuned to resonance. Under these circumstances the r-f current divides equally through C_{PG} and C_N. The resulting r-f currents in the plate tank flow in opposite directions and cancel the tank inductive effect so that no resonant build-up occurs between the coil and capacitors. A neon glow bulb, a loop of wire attached to a small flashlight bulb, or a sensitive r-f galvanometer may be used if an oscilloscope is not available.

If there is a milliammeter in the amplifier grid circuit, the adjustment of C_N may be made by observing the grid meter as the plate tank is tuned through resonance, with no plate voltage applied. When there is an unbalance between C_{PG} and C_N, the plate becomes alternately positive and negative as the plate tank strikes resonance. On positive swings, plate current flows. As the plate tank circuit is tuned to the resonant frequency, some of the electrons that were going to the grid now go to the plate, thereby causing a dip in grid current.

However, as C_N is adjusted to neutralize the amplifier stage, the r-f current from the input stage divides equally and flows in opposite directions in the two halves of the plate-tank coil, thus canceling the inductive effect of the coil and preventing the build-up of resonance in the tank. There is no rise in tank current and voltage, and the triode plate remains at zero potential. Therefore, with C_N properly adjusted, no dip in grid current occurs as the plate tank is tuned through the resonant frequency.

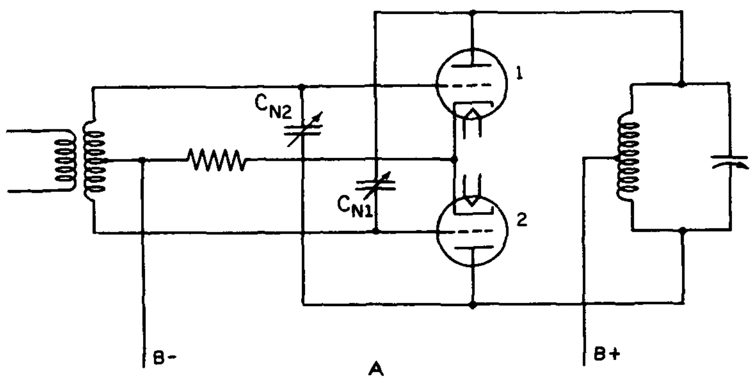
Another indication of the neutralized condition of the amplifier stage is obtained by observing the reaction on the plate and grid currents of the INPUT stage as the amplifier plate tank (with no voltage applied) is tuned through resonance. If the amplifier is not properly neutralized, the resonant build-up in the plate tank varies the load on the exciter stage as the plate tank is tuned through resonance. Thus, in the exciter stage, the grid current decreases and the plate current increases as the amplifier plate tank strikes resonance. However, when C_N is properly adjusted to neutralize the amplifier stage, no increase in loading occurs on the exciter stage as the plate tank is tuned to the resonant frequency, and the grid and plate currents in the exciter stage remain the same.

In some transmitter circuits it is more convenient to turn off the filament voltage on the amplifier stage instead of removing plate voltage. If this is done, the process of neutralizing the amplifier is carried out the same way, except that no current flows in the amplifier grid circuit. The absence of radio frequency in the amplifier plate tank, as evidence of the correct adjustment of C_N, may be determined by the effect on the exciter stage or on an r-f pick-up coil and associated indicator, as previously mentioned.

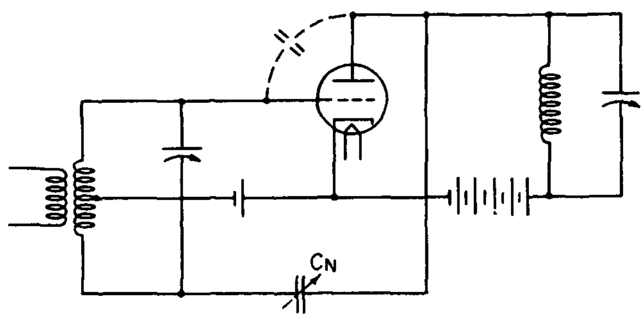
To obtain complete neutralization the transmitter must be designed so that there is no coupling between the input (grid) and output (plate) circuits of the amplifier stages other than through the interelectrode capacitance of the tubes. The input and output inductors must be shielded from each other or mounted at right angles to reduce any coupling between

them to a negligible amount. The wiring and arrangement of component parts must reduce stray capacitive or inductive coupling to a minimum.

Cross neutralization of a push-pull amplifier is accomplished as shown in figure 9–11, A. The plate of tube 1 is



A
PUSH-PULL NEUTRALIZATION



B
RICE SYSTEM OF NEUTRALIZATION

Figure 9–11.—Neutralization circuit.

connected to the grid of tube 2 through neutralizing capacitor C_{N1}, and the plate of tube 2 is connected to the grid of tube 1 through neutralizing capacitor C_{N2}. The voltage fed back through the interelectrode capacitance (plate-to-grid) of

tube 1 to the input circuit for that tube is counteracted by the voltage fed back through C_{N2}. The voltage fed back through the interelectrode capacitance (plate-to-grid) of tube 2 to the input circuit for tube 2 is counteracted by the voltage fed back through neutralizing capacitor C_{N1}.

A special method of amplifier neutralization, known as the Rice system, is shown in figure 9–11, B. This arrangement is similar to that of figure 9–11, A, except that the Rice system utilizes a split input circuit in place of a split output circuit. The voltage fed back from the plate to the grid through C_{PG} is counteracted by the voltage fed back through C_N. This circuit is a form of grid neutralization.

The use of a well-shielded tetrode or pentode makes neutralization unnecessary, because the plate and grid are shielded from each other by the screen grid and its associated r-f bypass capacitor which holds the screen at r-f ground potential. However, the over-all efficiency of these tubes is not as great as that of triodes, since there is a screen-grid power loss. The high impedance of such tubes makes them more suitable for voltage amplifiers than for final output stages where power output is the principal factor. Low excitation requirements make tetrodes and pentodes especially suitable for use in the intermediate stages of a transmitter.

Parasitic Oscillations

Circuit conditions in an oscillator or amplifier may be such that secondary oscillations occur at frequencies other than that desired. The frequency of these oscillations is neither that of the fundamental nor its harmonics. Such oscillations are appropriately termed PARASITIC OSCILLATIONS and are to be avoided. The energy required to maintain parasitic oscillations is wasted so far as useful output is concerned. A circuit afflicted with parasitics has low efficiency and frequently operates erratically.

Figure 9–12 shows some of the incidental circuits that may give rise to parasitics in a transmitter amplifier. The dotted lines in figure 9–12, A, outline a possible h-f circuit,

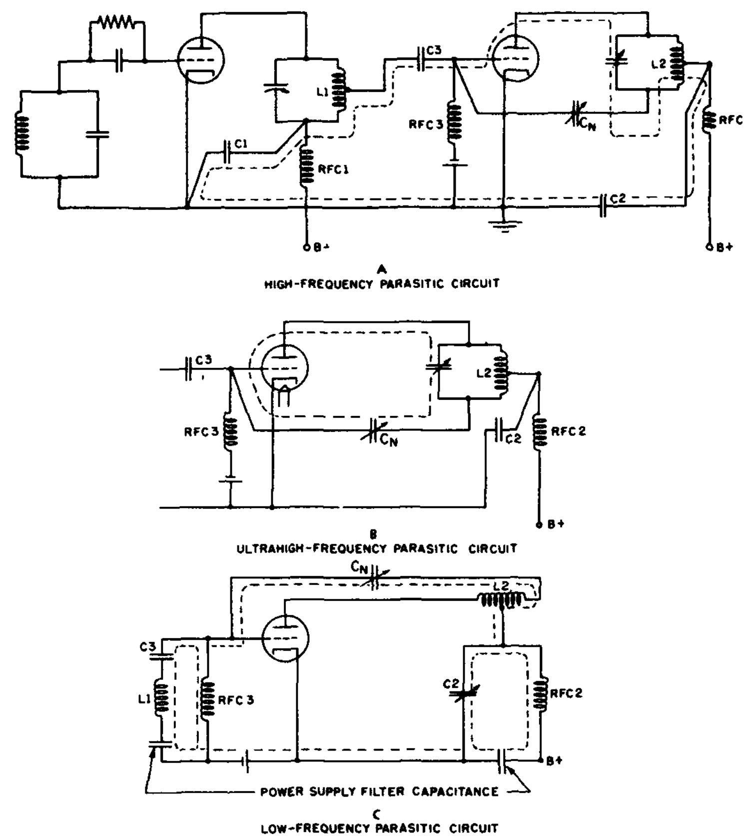


Figure 9-12.—Parasitic oscillatory circuits in a transmitter.

and those in figure 9–12, B, outline a possible u-h-f circuit. The part of a transmitter that constitutes a possible l-f parasitic circuit is shown in figure 9–12, C.

Parasitic oscillations may be suppressed by placing resistors or chokes at appropriate positions in the circuits, or by slightly modifying the existing values of certain circuit elements. Also, care should be used in the physical arrangement and wiring of parts. Parasitic suppressors consisting

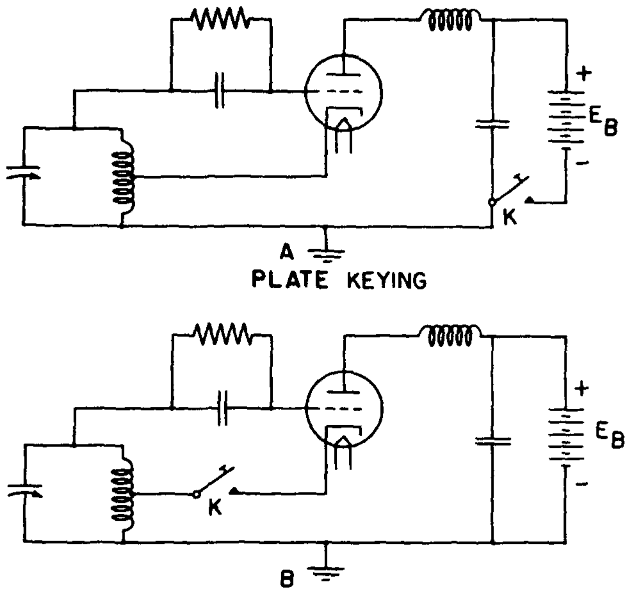
of an inductor and resistor in parallel are sometimes inserted in the grid and plate leads of an r-f amplifier to suppress high-frequency parasitic oscillations. The resistor has a resistance of from 50 to 100 ohms. The path through the r-f choke has a low impedance to mormal frequencies and a high impedance to high frequencies (parasitics). Thus, normal frequency currents flow through the r-f choke without attenuation. The path of the h-f parasitic currents is through the resistor which dissipates the feedback energy in the form of heat and reduces the magnitude of the parasitic oscillations to a negligible amount.

Keying Systems

Keying a c-w transmitter causes an r-f signal to be radiated ONLY when the key contacts are closed. When the key is open the transmitter does not radiate energy. Keying is accomplished in either the oscillator or amplifier stages of a transmitter. A number of different keying systems are used in Navy transmitters.

In most Navy transmitters the hand telegraph key is at low potential with respect to ground. The keying bar is usually grounded to protect the operator. Generally a keying relay with its contacts in the center-tap lead of the filament transformer is used to key the equipment. Because one or more stages use the same filament transformer, these stages are also keyed. The final amplifier, which is operated class C, is usually not keyed because with no excitation applied no space current flows. Hence, keying the final amplifier along with the other stages is not necessary.

Two methods of oscillator keying are shown in figure 9–13. In figure 9–13, A, the grid circuit is closed at all times, and the key opens and closes the negative side of the plate circuit. This system is called PLATE KEYING. When the key is open, no plate current can flow and the circuit does not oscillate. In figure 9–13, B, the grid and plate circuits are both open when the key is open and both are closed when the key is closed. This system is called CATHODE

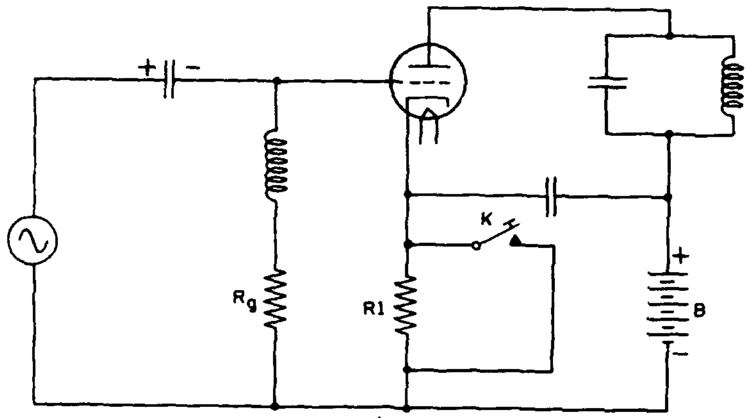


CATHODE KEYING

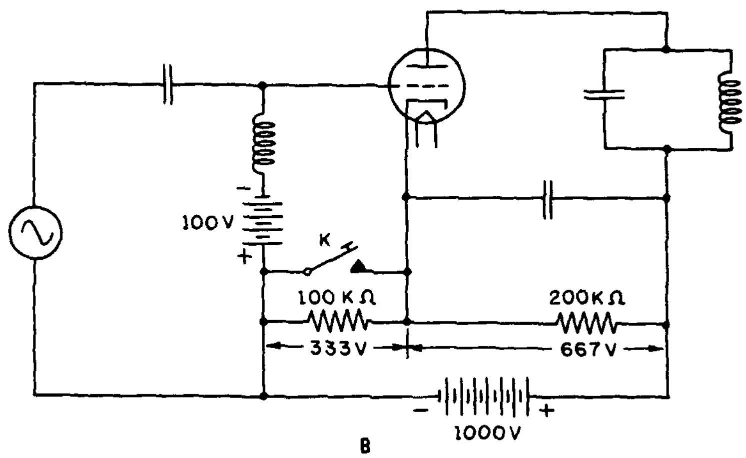
Figure 9–13.—Oscillator keying.

KEYING. Although the circuits of figure 9–13 may be used to key amplifiers, other keying methods are generally employed because of the larger values of plate current and voltage encountered.

Two methods of blocked-grid keying are shown in figure 9–14. The key in figure 9–14, A, shorts cathode resistor $R1$, allowing normal plate current to flow. With the key open, reduced plate current flows up through resistor $R1$, making the end connected to grid resistor R_g negative. If $R1$ has a high enough value, the bias developed is sufficient to cause practical cutoff of plate current. Complete cutoff is not possible because the bias voltage developed across $R1$ depends on the flow of some plate current through it. However, the blocking is sufficient for practical keying. Depressing the key short-circuits $R1$, thus increasing the bias above cutoff



A
KEY ACROSS CATHODE RESISTOR



B
KEY ACROSS GRID RESISTOR

Figure 9–14.—Blocked-grid keying.

and allowing the normal flow of plate current. Grid resistor R_g is the usual grid-leak resistor for normal bias. This method of keying is applied to the buffer stage in a c-w transmitter.

The blocked-grid keying method shown in figure 9–14, B, affords complete cutoff of plate current and is one of the best methods for keying amplifier stages in c-w transmitters. In the voltage divider, with the key open, two-thirds of 1,000 volts, or 667 volts, are developed across the 200 k-ohm resistor and one-third of 1,000 volts, or 333 volts, are developed across the 100 k-ohm resistor. The grid bias is $-100 -333$, or -433 volts. Because this is below cutoff, no plate current flows. The plate voltage is 667 volts. With the key closed, the 100 k-ohm resistor is shorted out and the voltage across the 200 k-ohm resistor is increased to 1,000 volts. Thus, the plate voltage becomes 1,000 volts at the same time the grid bias becomes -100 volts. Grid bias is now above cutoff and the amplifier triode conducts. Normal amplifier action follows.

Where greater frequency stability is required, the oscillator should remain in operation continuously while the transmitter is in use. This procedure keeps the oscillator tube at normal operating temperature and offers less chance for frequency variation to occur each time the key is closed. If the oscillator is to operate continuously and the keying is to be accomplished in an amplifier stage following the oscillator, the oscillator circuit must be carefully shielded to prevent radiation and interference to the operator while he is receiving.

In transmitters using a crystal-controlled oscillator the keying is almost always in a stage following the oscillator. In the large transmitters (75 watts or higher) the ordinary hand key cannot accommodate the plate current without excessive arcing. Moreover, because of the high plate potentials used it is dangerous to operate a hand key in the plate circuit. A slight slip of the hand below the key knob might result in a bad shock; or, in the case of defective r-f plate chokes, a severe r-f burn might be incurred. In these larger transmitters, some local low-voltage supply, such as a battery or the filament supply to the transmitter, is used with the hand key to open and close a circuit through

the coils of a keying relay. The relay contacts in turn open and close the keying circuits of the amplifier tubes. A schematic diagram of a typical relay-operated keying system is shown in figure 9–15. The hand key closes the circuit from the low-voltage supply through coil L of the keying relay. The relay armature closes the relay contacts as a result of the magnetic pull exerted on the armature. The armature moves against the tension of a spring. When the hand key is opened, the relay coil is deenergized and the spring opens the relay contacts.

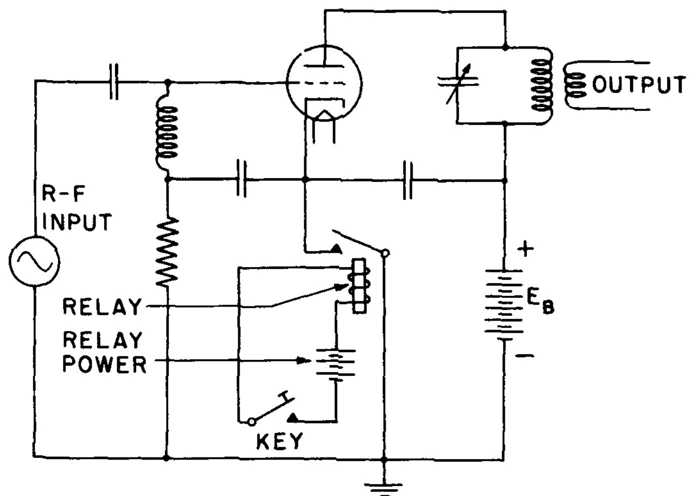


Figure 9–15.—Circuit for a relay-operated keying system.

Theoretically, keying a transmitter should instantly start and stop radiation of the carrier completely. However, the sudden application and removal of power creates large surges of current which cause interference in nearby receivers. Even though such receivers are tuned to frequencies far removed from that of the transmitter, interference is present in the form of clicks or thumps. To prevent such interference, key-click filters are used in the keying systems

of radio transmitters. Two types of key-click filters are shown in figure 9–16.

The capacitors and r-f chokes in both circuits of figure 9–16 prevent surges of current. The choke coil, *L*, causes a lag in the current when the key is closed, and the current builds up gradually instead of instantly. Capacitor *C*

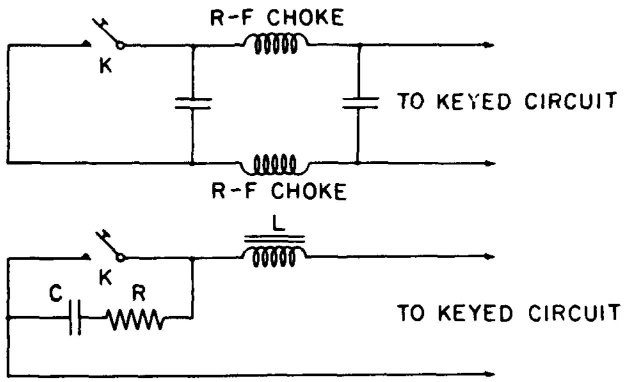


Figure 9–16.—Key-click filters.

charges up as the key is opened and slowly releases the energy stored in the inductor magnetic field. Resistor *R* controls the rate of charge and discharge of capacitor *C* and also prevents sparking at the key contacts by the sudden discharge of *C* when the key is closed.

Another difficulty that may be encountered in keying a transmitter is the presence of a back wave. A back wave results when some r-f energy leaks through to the antenna even though the key is open. The effect is as though the dots and dashes were simply louder portions of a continuous carrier. It may be difficult to distinguish the dots and dashes under such conditions. Back-wave radiation is usually the result of incomplete neutralization.

Circuit of a C-W Transmitter

The circuits of a small, shore-based, c-w transmitter and its power supply are shown in figure 9–17. The transmitter

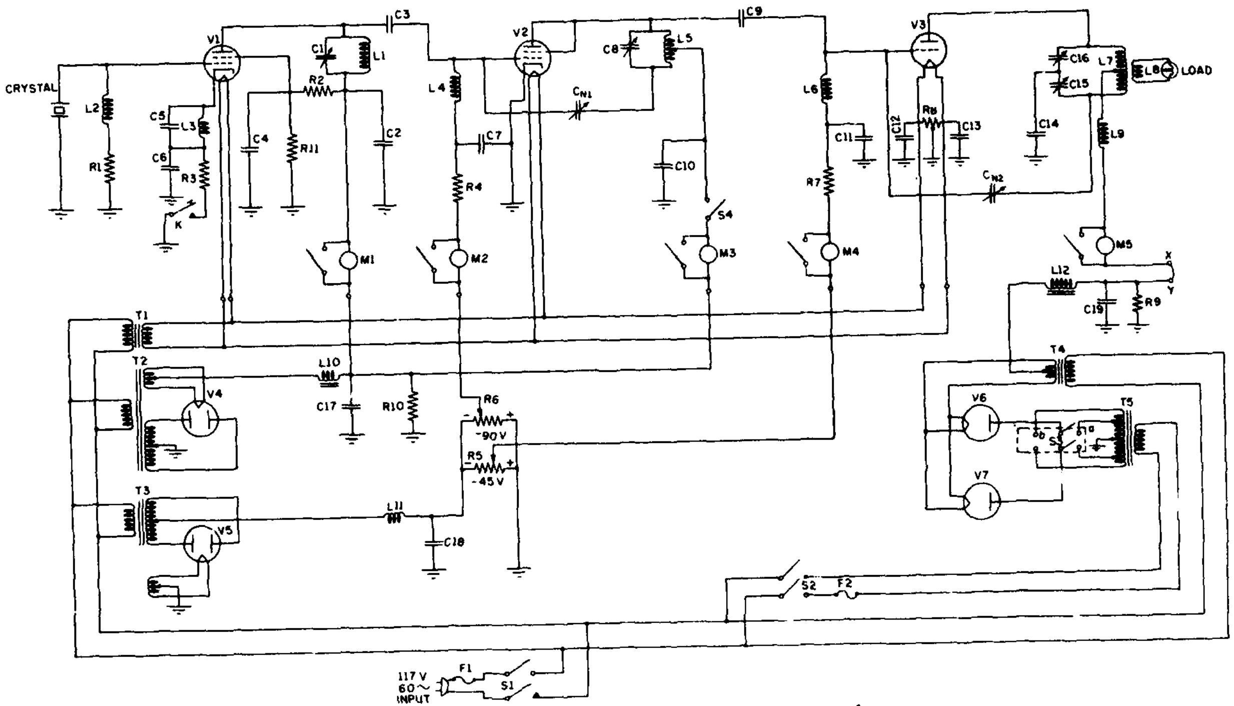


Figure 9-17.—C-w transmitter and power supply.

includes 3 tubes—(1) oscillator, (2) buffer, and (3) power amplifier. The power supply is designed to plug into a 117-volt 60-cycle source.

To simplify the wiring diagram, individual meters are shown in the plate and grid leads for each stage. In practice, one meter and a multi-terminal switch for the grid circuits and a similar combination for the plate circuits would suffice.

The crystal oscillator stage is keyed in the cathode circuit of beam-power tetrode *V*1, which is similar to a tuned-grid tuned-plate oscillator except that the crystal replaces the tuned-grid circuit. The stage oscillates with the key closed when the plate tank, *L*1*C*1, is tuned to a frequency close to that of the crystal. Grid-leak bias is developed across *R*1. The key-click filter circuit consists of *R*3, *L*3, *C*5, and *C*6.

The r-f excitation voltage appears across the r-f choke, *L*2. Screen voltage is applied through dropping resistor *R*2. Capacitor *C*4 is the screen bypass capacitor, which holds the screen at r-f ground potential. Capacitor *C*2 is the plate bypass capacitor, which places the lower end of the plate tank circuit at r-f ground. The output voltage across the tank is coupled to the buffer stage through capacitor *C*3. Meter *M*1 indicates the sum of the plate and screen currents in *V*1. The output power of the oscillator is about 5 watts at the crystal frequency.

The buffer stage, *V*2, employs a triode-connected beam-power tetrode biased for class-C operation and using plate neutralization. Neutralizing capacitor C_{N1} couples the correct amount of feedback voltage to the grid to neutralize the stage.

The oscillator output voltage is developed across the r-f choke, *L*4. Capacitor *C*7 is an r-f bypass capacitor that places the lower end of the choke at r-f ground potential. In addition to separate bias, the grid circuit uses the voltage developed across resistor *R*4 as automatic bias. Meter *M*2 indicates the buffer grid current. The stage may be used as a frequency doubler, in which case the plate tank tuning capacitor, *C*8, is adjusted so that the tank strikes resonance at the second harmonic of the crystal fundamental frequency.

Capacitor $C10$ is a plate bypass capacitor, which holds the center tap of tank coil $L5$ at r-f ground potential. Plate current is indicated by meter $M3$. The stage develops an output power of about 15 watts. The output voltage is coupled through capacitor $C9$ to the grid circuit of power amplifier triode $V3$. The excitation voltage appears across r-fh coke $L6$.

The power amplifier triode is biased for class-C operation and the plate tank ($L7$, $C15$, and $C16$) is tuned to the frequency of the grid-excitation voltage. The antenna is coupled to the final tank through link coil $L8$. Resistor $R8$ is center-tapped to provide a common return for the plate and grid circuits and to prevent the 60-cycle filament voltage from modulating the r-f grid voltage. Capacitor $C11$ is an r-f bypass capacitor that holds the lower end of r-f choke $L6$ at r-f ground potential. Capacitors $C12$ and $C13$ are filament bypass capacitors that keep r-f current out of the filament leads. Capacitor $C14$ effectively places the center tap of plate tank coil $L7$ at r-f ground potential. The stage employs plate neutralization. The neutralizing capacitor, C_{N2}, couples a portion of the voltage between the lower end of the plate tank and ground back to the grid, to neutralize the voltage fed back through the plate-to-grid capacitance of $V3$. The r-f choke, $L9$, keeps r-f currents out of the plate supply lead. Meter $M4$ indicates grid current and $M5$ plate current. Power output is approximately 100 watts when the plate supply voltage is 1,000 volts.

The power supply includes three full-wave rectifiers to provide plate, screen, and grid voltages for the transmitter tubes. It also includes filament supply transformers for both transmitter and power supply tubes. Transformer $T1$ supplies filament power to $V1$, $V2$, and $V3$.

Switches $S1$ and $S2$ are arranged so that $S1$ must be closed before $S2$ can be energized. Thus, filament and bias voltages are provided for $V2$ and $V3$ before plate voltage can be applied to $V3$. The primaries of all the power supply transformers are connected in parallel and are supplied by a 117-volt 60-cycle source. Power supplies are treated in chapter 3.

All radio transmitters must be properly tuned to ensure efficient operation on the assigned frequency. Transmitters are always tuned on c-w even if m-c-w or voice modulation may also be used. Plate-current meters are generally used to indicate proper adjustment of the r-f stages. All stages, with the exception of the oscillator, are always adjusted or tuned for minimum plate current. If a stage is not tuned to resonance, the plate current will be high and high plate dissipation, power loss, and low output will result. When a stage is loaded by another stage or an antenna, the plate current of the stage in question must be rechecked for circuit resonance (minimum plate current) after loading.

If a gassy tube is present in the set, plate current in that stage cannot be brought to the proper minimum, and grid current will remain too low or may even reverse. The tube will act as though there were a short between the grid and cathode and much of the energy supplied to the stage will be grounded and lost. This condition can be recognized by any of the indications just mentioned and by a violet-colored glow between the tube elements. The only remedy for this condition is a new tube.

The coupling of the tuned antenna to the transmitter is accompanied by three principal effects—

1. The antenna current (r-f energy) increases.
2. The plate current of *V*3, as indicated on meter *M*5, increases.
3. The grid current of *V*3, as indicated on meter *M*4, decreases.

In the final tune-up process the act of moving the antenna link coil closer to the final power-amplifier tank coil usually detunes the final stage slightly. This detuning results in an increase in the indication on plate-current meter *M*5. To correct this condition, plate tank capacitors *C*15 and *C*16 should be readjusted until minimum current is indicated on *M*5. This adjustment results in a further increase in output power and antenna current. The antenna is then ready for keying.

Capabilities of a C-W Transmitter

In view of the comparative slowness and inconvenience of keying the dots and dashes of Morse code, it might seem that radiotelegraphy would be superseded by radiotelephony, which uses modulated waves. C-w transmission, however, has four distinct advantages over radiotelephony.

1. Radiotelegraph transmitters have a greater transmission range than radiotelephone transmitters of the same power output because, in the latter, speech from a distant point may be audible, but not intelligible.
2. C-w signals may be picked up by code receivers that are capable of rejecting most of the interference characteristic of all r-f waves.
3. The comparable radiotelegraph transmitter is smaller and much simpler to operate.
4. Within a given frequency band, many more radiotelegraph transmitters than radiotelephone transmitters may be operated without interference.

AMPLITUDE-MODULATED RADIOTELEPHONE TRANSMITTER

Amplitude modulation has been defined as the variations of the magnitude of the r-f output of a transmitter at an audio rate. In other words, the r-f energy increases and decreases in accordance with the energy delivered by the audio modulator. If the audio frequency is high, the radio frequency varies in amplitude more rapidly than if the audio frequency were low. If the audio note is loud in volume, the r-f energy is increased and decreased by a larger percentage than if the audio note were soft. Thus, the r-f variations correspond with the a-f variations.

A microphone or a similar device is used to produce the electrical equivalent of the audio signal. The signal is then amplified by means of an a-f amplifier before it is fed to the modulator. Because of the importance of microphones in the communications chain a brief description of some of the

more common microphones, together with their characteristics, follows.

Microphones

A microphone is essentially an energy converter that changes acoustical (sound) energy into corresponding electrical energy. When one speaks into a microphone, the audio pressure waves strike the diaphragm of the microphone and cause it to move in and out in accordance with the instantaneous pressure delivered to it. The diaphragm is attached to a device that causes current to flow in proportion to the instantaneous pressure applied to the diaphragm. Many devices can perform this function, each having characteristics that make its use advantageous under a given set of circumstances.

Most microphones, with the exception of the carbon microphone, are relatively inefficient—that is, the output in electrical energy is considerably less than the input in acoustical energy. Some, however, are more efficient than others; and some have a better frequency response than others. The characteristics of microphones, therefore, will be discussed before the various types of microphones are discussed. Microphones are rated according to their (1) frequency response, (2) impedance, and (3) sensitivity.

Frequency response.—For good quality, the electrical waves from a microphone must correspond closely in magnitude and frequency to the sound waves that cause them, so that no new frequencies are introduced. The frequency range of the microphone (that range of frequencies over which the microphone is capable of responding) must be no wider than the desired over-all response limits of the system with which it is to be used. The microphone response should be uniform, or flat, within its frequency range and free from sharp peaks or dips such as those caused by mechanical resonances. To aid in attaining this condition, some form of damping may be employed.

Impedance.—Crystal microphones have impedances of several hundred thousand ohms; whereas magnetic and

dynamic microphones have impedances that range from 20 to 600 ohms. The impedance of a microphone is usually measured between its terminals when some arbitrary frequency in the useful audio range—for example, 1,000 cycles—is used.

The impedance of magnetic and dynamic microphones varies with frequency in much the same manner as that of any coil or inductance—that is, the impedance rises with increasing frequency. The actual impedance of a microphone is of importance chiefly as it is related to the load impedance into which the microphone is designed to operate. If the load has a high impedance, the microphone should have a high impedance, and vice versa. Of course, impedance-matching devices may be used between the microphone and its load.

A long transmission line between the microphone and the amplifier input tends to seriously attenuate the high frequencies, especially if the impedance of the microphone is high. This action results from the increased capacitive effect of the line at the higher frequencies. If the microphone has a high impedance the high-frequency currents drawn through the inherent capacity of the line cause an increased voltage loss within the microphone, and therefore less voltage is available at the load. Because the voltage generated by the microphone is very minute, all losses in the microphone and the line must be kept to a minimum. At the lower frequencies the capacitive effect is less and the losses are correspondingly less. If the microphone has a low impedance a correspondingly lower voltage drop will occur in the microphone and more voltage will be available at the load.

Because many microphone lines aboard ship are long, it is necessary to use low-impedance microphones in order to preserve a satisfactory signal voltage level over the required audio band at the input grid of the amplifier.

Sensitivity.—The sensitivity or efficiency of a microphone is usually expressed in terms of the electrical power level which the microphone delivers to a terminating load (the impedance of which is equal to the rated impedance of the

microphone) compared to the acoustical intensity level or pressure of the sound energy that is being picked up. Because sound energy at the input is being compared with electrical energy at the output, some basis of comparison must be established.

One method is to assume that a microphone has a sensitivity of 0 db (the level of comparison) if a force of 1 dyne per square centimeter on the diaphragm produces an output of 1 volt on open circuit. The pressure of 1 dyne per square centimeter was chosen because that is the approximate pressure produced by normal speech on the diaphragm of a microphone held a few inches from the mouth. The usual method, however, is to assume that the 0 db level represents an input of 1 dyne per square centimeter (as in the first method) and an output of 1 milliwatt. If it is further assumed that the milliwatt is developed in 600 ohms, then dbm or volume units (vu) may be used. Decibels and the various power-level units are given in chapter 6; a power-level chart is shown in figure 6–11.

Suppose a microphone is rated at -80 db. This rating means that the energy output is much less than the energy input. Actually, the output is 10^{-8} milliwatt for an equivalent input of 1 milliwatt, and this is equivalent to -80 db. This rating may be demonstrated by the use of equation 6–4 in chapter 6—

$$\text{db} = 10 \log_{10} \frac{P_2}{P_1}$$

$$\text{db} = 10 \log_{10} \frac{(10)^{-8}}{1} = -80 \text{ db}.$$

It is important to have the sensitivity of the microphone as high as possible. High sensitivity means a high electrical power output level for a given input sound level. High microphone output levels require less gain in the amplifiers used with them and thus provide a greater margin over thermal noise, amplifier hum, and noise pick-up in the line between the microphone and the amplifier.

When a microphone must be used in a noisy location, an additional desirable characteristic is the ability of the microphone to favor sounds coming from a nearby source over random sounds coming from a relatively greater distance. Microphones of this type tend to cancel out random sounds and to pick up only those sounds originating a short distance away. When talking into this type of microphone the lips must be held as close as possible to the diaphragm. Directional characteristics also aid in discriminating against background noise.

CARBON MICROPHONE.—The carbon microphone is the most common type of microphone. It operates on the principle that a change in sound pressure on a diaphragm that is coupled to a small volume of carbon granules will cause a corresponding change in the electrical resistance of the granules.

The single-button carbon microphone (fig. 9–18, A) consists of a diaphragm mounted against carbon granules that are contained in a small cup. In order to produce an output voltage, this microphone is connected in a series circuit containing a battery and the primary of a microphone transformer. The pressure of the sound waves on the diaphragm, which is coupled to the carbon granules, causes the resistance of the granules to vary. Thus a varying direct current in the primary produces an alternating voltage in the secondary of the transformer. This voltage has essentially the same waveform as that of the sound waves striking the diaphragm. The current through a carbon microphone may be as great as 0.1 ampere, and the resistance may vary from about 50 to 90 ohms. The voltage developed across the secondary depends upon the turns ratio and also upon the rate of change in primary current. Normal output voltage of a typical circuit is from 3 to 10 volts peak across the secondary terminals.

The double-button carbon microphone is shown schematically in figure 9–18, B. Here one button is positioned on each side of the diaphragm so that an increase in pressure and resistance on one side is accompanied simultaneously by

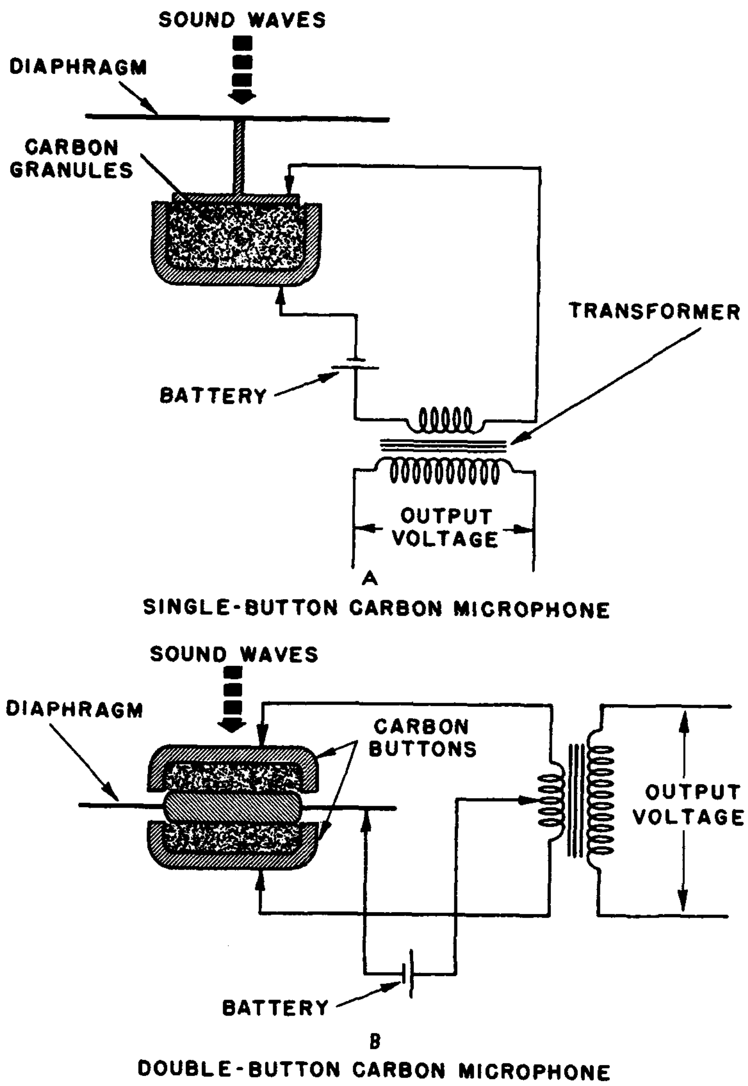


Figure 9-18.—Schematic diagram of carbon microphones.

a decrease in pressure and resistance on the other. Each button is in series with the battery and one half the transformer primary. The decreasing current in one half of the primary and the increasing current in the other half produce an output voltage in the secondary that is proportional to the sum of the primary signal components. This action is similar to that of push-pull amplifiers (chapter 6).

Commercial types of carbon microphones give essentially faithful reproduction from 60 to 6,000 cycles, and their output is of the order of -50 db.

The carbon microphone has the disadvantage of requiring an external voltage source; it may be noisy; and unless the necessary precautions are taken in the design the microphone tends to peak up (have mechanical resonance) at certain frequencies.

Dynamic microphone.—The dynamic, or moving-coil, microphone (fig. 9-19) consists of a coil of wire attached to a diaphragm and is so constructed that the coil is suspended and free to move in a radial magnetic field. Sound waves

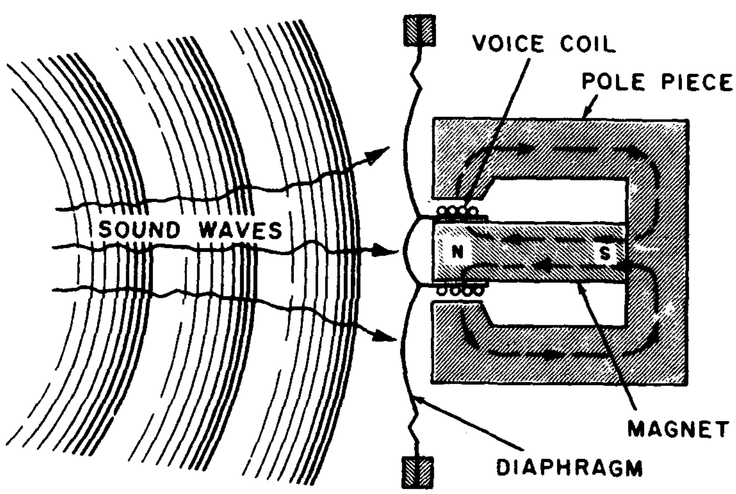


Figure 9-19.—Action of a dynamic microphone.

impinging on the diaphragm cause the diaphragm to vibrate. This vibration moves the voice coil through the magnetic field so that the turns cut the lines of force in the field. This action generates a voltage in the coil that has the same waveform as the sound waves striking the diaphragm.

The dynamic microphone requires no external voltage source, has good fidelity (approx. 20 to 9,000 cycles with proper damping), is directional for high-frequency sounds, and has an output of the order of -85 db. The impedance of the dynamic microphone is low (50 ohms or less). Therefore, it may be connected to relatively long transmission lines without excessive attenuation of the high frequencies.

CRYSTAL MICROPHONE.—The crystal microphone utilizes a property of certain crystals—such as quartz and Rochelle salt—known as the PIEZOELECTRIC EFFECT, treated in chapter 7. The bending of the crystal, resulting from the pressure of the sound wave, produces an emf across the faces of the crystal. This emf is applied to the input of an amplifier.

The crystal microphone (fig. 9–20) consists of a diaphragm that may be cemented directly on one surface of the crystal (fig. 9–20, A), or in some cases it may be connected to the crystal element through a coupling member (fig. 9–20, B). A metal plate, or electrode, is attached to the other surface of the crystal. When sound waves strike the diaphragm, the vibrations of the diaphragm produce a varying pressure on the surface of the crystal, and therefore an emf is induced across the electrodes. This emf has essentially the same waveform as that of the sound waves striking the diaphragm.

Rochelle salt is most commonly used in crystal microphones because of its relatively high voltage output. Schematic diagrams showing how crystal microphones function are given in figure 9–20.

Actually, a large percentage of crystal microphones employ some form of the bimorph cell. In this type of cell two crystals, so cut and oriented that their voltages will be additive in the output, are cemented together and used in place of the single crystal.

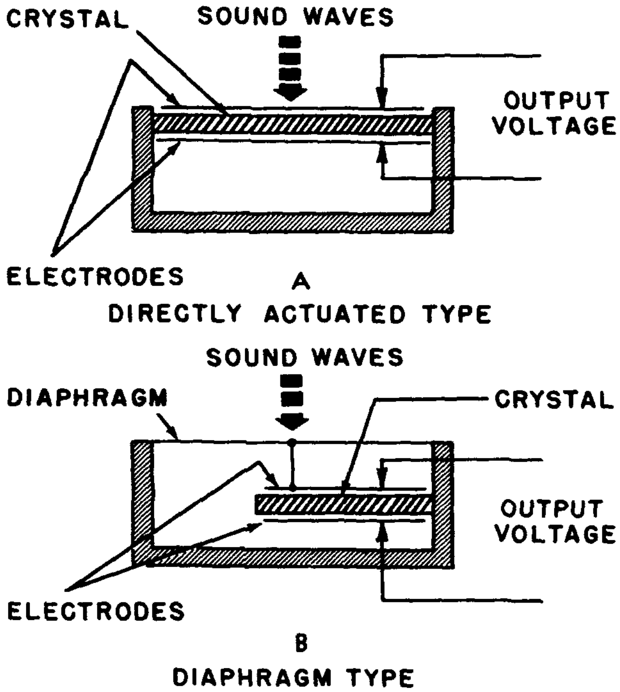


Figure 9-20.—Schematic diagrams of crystal microphones.

This type of microphone has high impedance (several hundred thousand ohms), is light in weight, requires no battery, is nondirectional, has good frequency response (up to 17,000 cps for the directly actuated type and between 80 to 6,000 cps for the diaphragm type), and has an output of the order of -70 db. However, the crystal microphone is sensitive to high termperature, humidity, and rough handling and therefore its use is restricted where these conditions prevail. Nevertheless, it is used extensively in broadcast work where its relatively high output is an advantage.

MAGNETIC MICROPHONE.—The magnetic, or moving-armature, microphone (fig. 9-21) consists of a permanent magnet and a coil of wire enclosing a small armature. Sound

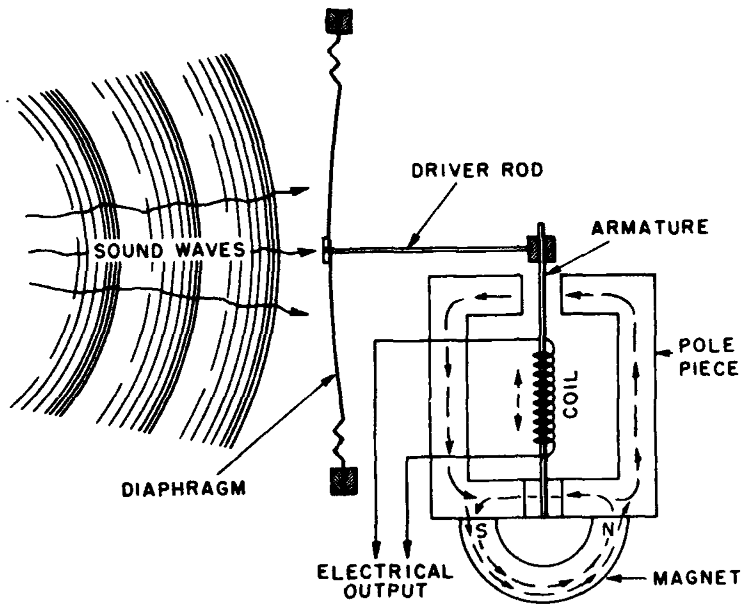


Figure 9-21.—Action of a magnetic microphone.

waves impinging on the diaphragm cause the diaphragm to vibrate. This vibration is transmitted through the drive rod to the armature, which vibrates in a magnetic field, thus changing the magnetic flux through the armature and consequently through the coil.

When the armature is in its normal position midway between the two poles, the magnetic flux is established across the air gap, and there is no resultant flux in the armature.

When a compression wave strikes the diaphragm, the armature is deflected to the right. Although a considerable amount of the flux continues to move in the direction of the arrows, some of it now flows from the north pole of the magnet across the reduced gap at the upper right, down through the armature, and around to the south pole of the magnet. The amount of flux flowing down the left-hand pole piece is reduced by this amount.

When a rarefaction wave strikes the diaphragm, the armature is deflected to the left. Some of the flux is now directed from the north pole of the magnet, up through the armature, through the reduced gap at the upper left, and back to the south pole. The amount of flux now moving up through the right-hand pole piece is reduced by this amount.

Thus, the vibrations of the diaphragm cause an alternating flux in the armature. The alternating flux cuts the stationary coil wound around the armature and induces an alternating voltage in the coil. This voltage has essentially the same waveform as that of the sound waves striking the diaphragm.

The magnetic microphone is the type most widely used in shipboard announcing and communicating systems because it is more resistant to vibration, shock, and rough handling than other types of microphones.

There are other types of microphones, such as the ribbon velocity microphone and the capacitor microphone, that are treated in the rating texts. All of the foregoing microphones may be used for radiotelephone broadcast, but the circumstances usually limit the choice to one or two types.

Circuits of an A-M Radiotelephone Transmitter

The block diagram of a simple a-m radiotelephone transmitter is shown in figure 9–22. The oscillator, buffer stage,

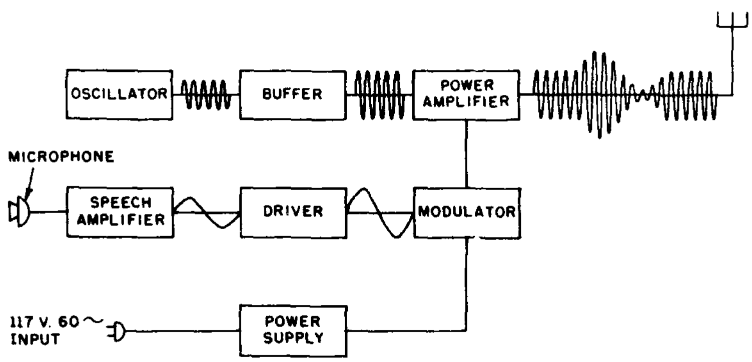


Figure 9–22.—Block diagram of an a-m radiotelephone transmitter.

and power amplifier closely resemble the c-w transmitter.

In figure 9–17 the c-w transmitter is keyed by opening and closing the cathode circuit of oscillator $V1$, which turns the r-f output of $V2$ and $V3$ off and on. If it is desired to vary the output of the transmitter instead of merely turning it off and on, the voltage on one of the elements of the final r-f power-amplifier tube, $V3$, may be varied. For example, if the plate voltage of $V3$ were varied at the audio-frequency rate, the output of the amplifier, and hence of the transmitter, would be varied at the same rate. This method, known as PLATE MODULATION, is used in the following example and is the most popular type of amplitude modulation. Plate modulation is discussed in greater detail in chapter 8.

In order to vary the plate voltage of the final r-f amplifier it is necessary first to produce an audio voltage. An audio voltage may be produced with a microphone. The output of a microphone is, however, very low (usually less than 1 volt), while the plate voltage of the r-f amplifier is quite high. The insertion of a small audio voltage in series with a high plate voltage would result in only a small variation of the plate voltage.

It is necessary, therefore, to amplify the electrical output of the microphone before it is applied in series with the plate of the final r-f power amplifier. This amplification is accomplished in three units of the block diagram of figure 9–22. These units are the speech amplifier, the driver, and the modulator. The output of the microphone is fed to the grid of $V1$, a class-A voltage amplifier pentode (fig. 9–23), which is the input tube of a two-tube speech amplifier. The output of the speech amplifier supplies the driver unit ($V3$), which further amplifies the audio signal to drive the audio modulator tubes ($V4$ and $V5$). The output voltage of the modulator tubes is fed in series with the plate supply voltage of the final r-f power amplifier of the transmitter. The modulator can be any type of audio power amplifier capable of providing sufficient undistorted power. Thus, it may be a class-A, class-AB, or class-B amplifier. In this example it is a class-AB push-pull stage.

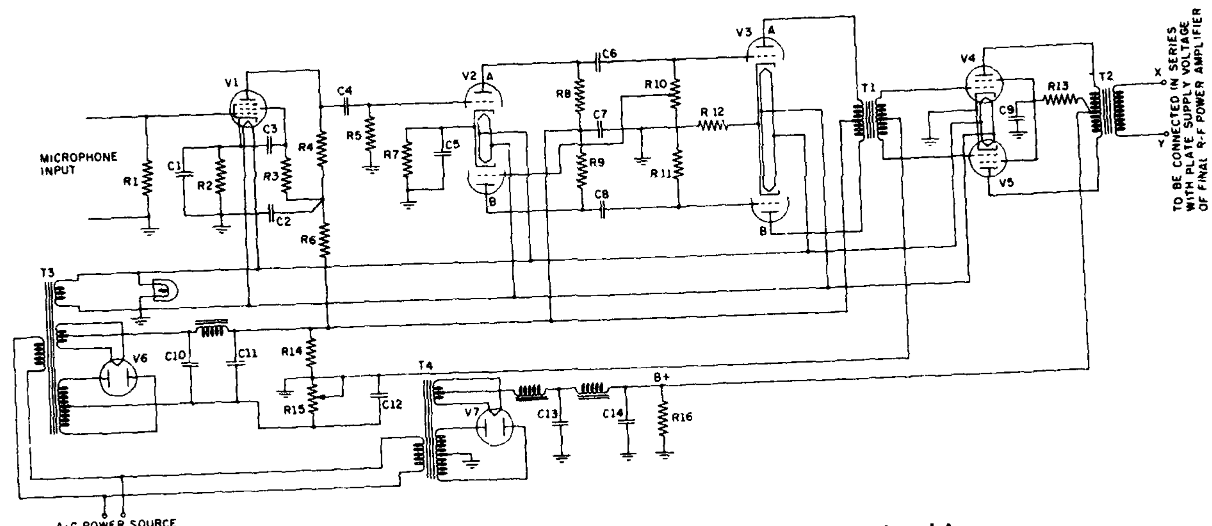


Figure 9-23.—Schematic diagram of a speech amplifier, driver, and modulator.

The power supply unit includes the necessary transformers, rectifiers, and filters, to supply the filaments, plates, screens, and grid-bias voltages from a single-phase 117-volt 60-cycle source.

The output of the modulator unit may be applied in series with the plate of the final r-f power amplifier (shown in fig. 9-17 below *M*5).

In the absence of a modulating signal, a continuous r-f wave is radiated by the antenna. Assume that an audio voltage of sine waveform is applied across *R*1 to the grid of *V*1. The amplified signal appears across plate resistor *R*4 and is coupled to the grid of *V*2*A* through *C*4. The amplified signal is applied to the grid of driver *V*3*A* through *C*6, and a part of it is applied to the phase-inverter stage, *V*2*B*, by means of the tap on *R*10. The amplified signal from *V*2*B* is simultaneously applied via *C*8 to the grid of *V*3*B* in opposite phase to that applied to the grid of *V*3*A*. The driver stage (*V*3*A* and *V*3*B*) provides excitation for the grids of modulator tubes *V*4 and *V*5 through impedance-matching transformer *T*1.

The push-pull amplifier (*V*4 and *V*5) develops a relatively large audio output voltage across the secondary of modulation transformer *T*2 to amplitude-modulate the carrier output 100 percent. The following three relationships exist in the case of 100-percent modulation.

1. The modulation is 100 percent when the peak value of the audio voltage across the secondary of transformer *T*2 is approximately equal to the plate supply voltage of *V*3 (fig. 9-17) that appears across resistor *R*9. The degree of modulation depends on the volume of sound striking the microphone, which in turn determines the magnitude of the audio signal voltage developed across *R*1, the grid input to *V*1 (fig. 9-23).
2. The modulation is 100 percent when the energy supplied to the final r-f amplifier by modulation transformer *T*2 is equal to one-half the r-f energy delivered to the final tank by the high-voltage power supply (*T*5, *V*6, and *V*7 in fig. 9-17). The mixing of the a-f voltage from *T*2

(fig. 9–23) with the r-f voltage developed across final tank coil $L7$ (fig. 9–17) produces side-band frequencies (sum and difference frequencies) that are coupled into the antenna circuit by the mutual inductance existing between $L7$ and coupling coil $L8$ of the antenna.

3. The modulation is 100 percent when the r-f energy delivered to the antenna (as a result of the injection of the a-f voltage from $T2$) is increased 50 percent above the amount delivered to the antenna when no audio signal is present. This condition represents an increase of 22.5 percent in the antenna current. Thus, the antenna r-f ammeter (not shown) may be used as a modulation indicator.

In this example there are four frequencies present in the final tank ($C15$, $C16$, and $L7$ of fig. 9–17). These are (1) the crystal frequency, (2) the audio frequency, (3) the sum of these two frequencies, and (4) the difference between these two frequencies. All except the audio frequency are coupled into the antenna circuit. The audio frequency is so far removed from the carrier and its associated side bands that that the mutual inductive coupling bewteen $L7$ and $L8$ for this frequecny is effectively zero.

The details of adjustment and operation of both a-m and f-m transmitters are included in instruction books on these equipments.

FREQUENCY-MODULATED RADIOTELEPHONE TRANSMITTER

Intelligence may also be conveyed by varying the frequency of a continuous radio wave of constant amplitude. The carrier frequency can be varied a small amount on either side of its average, or assigned, value by means of the a-f modulating signal. The amount the carrier is varied depends on the magnitude of the modulating signal, and the frequency with which the carrier is varied depends on the frequency of the modulating signal. The amplitude of the r-f carrier remains constant with or without modulation.

A radio receiver that is sensitive only to variations in the frequency of the incoming carrier and that discriminates to a large extent against variations in amplitude is used to receive the f-m signals.

Consider a typical f-m transmitter that is included in the main station equipment of an assembly designed for 2-way communications on the v-h-f band. The assembly includes a 50-watt f-m narrow-band transmitter, an f-m receiver, a power supply, and the necessary accessories for 2-way communications in a fixed station. However, for the purpose of this chapter only the transmitter portion of the equipment will be considered. Details of the various methods of obtaining frequency modulation are included in chapter 8.

Circuit of an F-M Radiotelephone Transmitter

The block diagram of a narrow-band f-m transmitter is shown in figure 9–24. All power from the unit is obtained

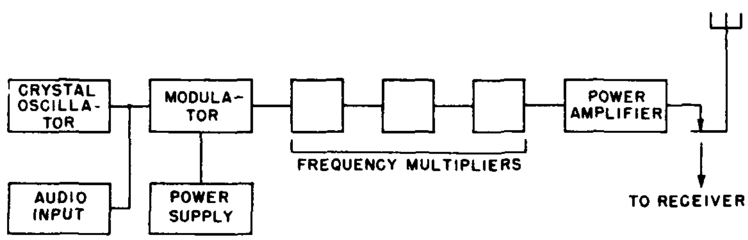


Figure 9–24.—Block diagram of a narrow-band f-m transmitter.

from a 115-volt 60-cycle source. The total drain is low enough so that the power may be taken from any convenient branch circuit outlet.

Oscillations are produced in the crystal-oscillator stage, the output of which is fed to a phase-shift network that supplies the grid voltage of the modulator tubes. The phase of the output voltage of the modulator varies in accordance with the input signal from the microphone. The phase shift is equivalent to a relatively low deviation of the output signal frequency of the modulator stage.

The frequency of the output of the modulator stage is quadrupled in the first multiplier stage, again quadrupled in the second multiplier stage, and doubled in the last multiplier the output of which drives the power-amplifier stage, which consists of two beam-power tubes in parallel. To obtain the final operating frequency, the crystal frequency is multiplied by 32.

A schematic drawing of the narrow-band f-m transmitter, shown in the block diagram in figure 9–24 is shown in figure 9–25. For convenience in making adjustments, a meter, $M1$, and meter switch, $S1$, are provided on the chassis to indicate the grid-circuit current of each stage.

With the exception of the first, the position numbers on the meter switch correspond with the numbers on the tops of the r-f transformers. For example, r-f transformer $T2$ is tuned for maximum current through the meter when the meter switch is in position 2. The transmitter employs the phase-shift method of obtaining frequency deviations. This method is discussed in chapter 8.

This transmitter exhibits characteristics that differ from those of the usual a-m type of transmitter. Intelligence is conveyed in the f-m transmitter by varying the frequency of the constant-amplitude carrier wave about an average assigned value. This process is in marked contrast with the amplitude-modulated transmitter previously described, in which intelligence is conveyed by varying the amplitude of the constant-frequency carrier wave.

The phase-shift method of obtaining frequency deviations in this f-m transmitter permits direct crystal control of the average carrier frequency. Frequency multiplication after modulation is necessary in order to generate the required frequency deviation of ± 15 kc on either side of the carrier.

Operation

Crystal oscillator $V1$ is a pentode connected as a triode and operated as a conventional triode crystal oscillator. The crystal is of the low-drift AT-cut type and operates at

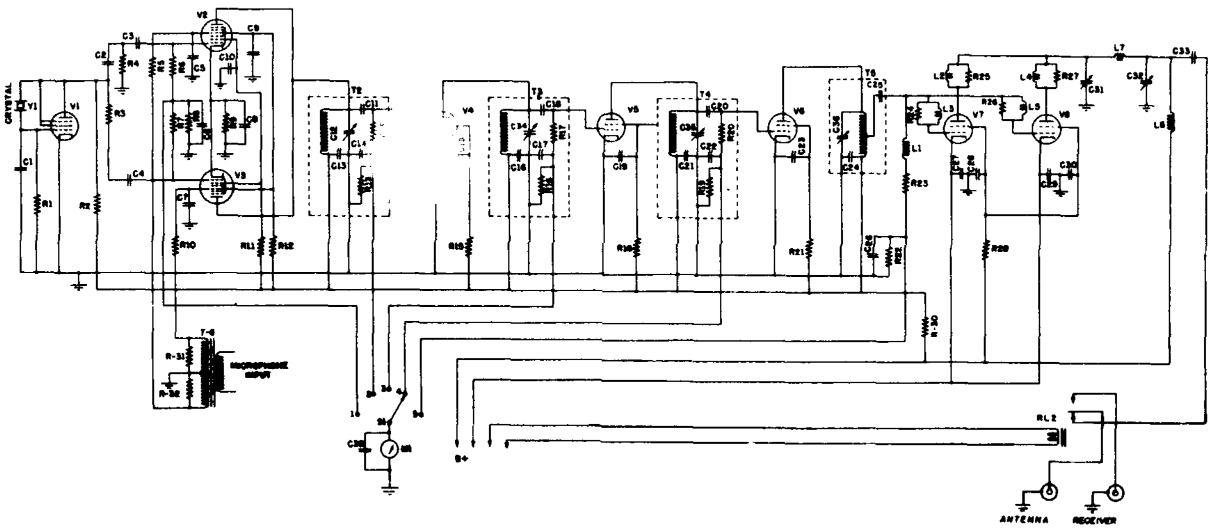


Figure 9-25.—Schematic diagram of a narrow-band f-m transmitter.

the 32nd subharmonic of the output frequency. The output-frequency range is from 30 to 40 megacycles, and thus the crystals that are used range in frequency from 937.5 kc to 1,250 kc. The crystal oscillator utilizes a resistance-coupled circuit so that no oscillator tuning is necessary when changing crystals. The crystal is connected between the grid and plate of *V*1, and *R*2 acts as the plate-circuit load.

The control grids of the two balanced modulators, *V*2 and *V*3, are driven from the plate of the r-f oscillator through a phase-shifting network that displaces their associated r-f driving voltages by 90°.

The plate currents of *V*2 and *V*3 are about 90° out of phase and equal in magnitude because the driving voltages from the oscillator are equal when there is no modulating signal on the number 4 grids of the tubes. The two plate currents add vectorially to produce a resultant current in *T*2 that is approximately 45° out of phase with each component. The output voltage of *T*2 varies in phase and magnitude with this resultant current.

When a modulating signal is applied to grid number 4 of *V*2 and *V*3, the plate currents are varied about the average values they would have if no modulation were present. For example, as grid number 4 of *V*2 swings in a positive direction the plate current of *V*2 increases. Simultaneously the voltage on grid number 4 of *V*3 swings in a negative direction, and the plate current of *V*3 decreases

Because these two currents are 90° out of phase, the resultant current (their vector sum) changes its phase with respect to the components as the components change in magnitude. Thus the current in *T*2, and the output voltage of *T*2, change in phase with the modulating signal. This change in phase is equivalent to a limited change in frequency occurring during the time that the output voltage phase shift is occurring.

The modulator grids of *V*2 and *V*3 are connected to the secondary of push-pull audio transformer *T*6. This transformer is driven directly from the microphone.

The modulator grids are fed through the frequency cor-

recting networks $R10C7$ and $R5C5$. These R-C combinations attenuate the a-f range (above 2,000 cycles) so that excessive frequency deviation is not obtained. Resistors $R31$ and $R32$ are terminating resistors for the secondary of microphone transformer $T6$. Cathode bias is obtained across $R9$ and $C8$. $R11$ and $R12$ are screen voltage dropping resistors and $C10$ and $C9$ their respective screen bypass capacitors.

The phase shift depends on the ratio of the signal strength of the carrier to the modulating signal strength. A ratio of 2 to 1 is equivalent to about 0.5 radian, or 30°, phase shift. The frequency shift is equal to the product of the modulating frequency and the phase-shift angle in radians.

FREQUENCY MULTIPLIERS.—The frequency deviation that may be produced by the balanced modulator stage, $V2$ and $V3$, is small—usually not more than half the modulating frequency, since the phase-shift angle is of the order of 0.5 radian. To get sufficient deviation (± 15 kc) the frequency of the modulated wave is multiplied by 32. As mentioned previously, this multiplication is accomplished by two quadruplers, $V4$ and $V5$, and a doubler, $V6$. All of these tubes act as class-C amplifiers with the plate tanks tuned to either the second or fourth harmonic of their respective grid signals. The grid drive in each case is such that plate current is well above saturation so that slight changes in tuning or reduction in tube emission have little effect on succeeding stages. All stages up to this point use receiving-type tubes working at relatively low plate and filament currents.

POWER AMPLIFIER.—The power amplifier utilizes two beam transmitting tubes, $V7$ and $V8$, in parallel as a class-C amplifier. Grid-leak bias is used; and, as in the previous stages, grid current is metered for alignment and testing. The plate tank and antenna circuit is of the pi-type for harmonic suppression and ease of adjustment. This circuit consists of plate tuning capacitor $C31$, tank coil $L7$, and antenna loading capacitor $C32$. The output is fed through blocking capacitor $C33$ to antenna relay $RL2$.

QUIZ

1. For a given transmitter, why does c-w have a greater range than m-c-w or voice modulations?
2. Why is the very-low-frequency band not covered by shipboard transmitters?
3. What is an advantage of using the very-low-frequency band?
4. Which frequency band is used for radar?
5. What is a disadvantage of using crystal-controlled oscillators in Navy transmitters operating at the lower frequencies?
6. How is frequency drift eliminated in Navy transmitters?
7. What type of oscillator is commonly used in the lower frequency ranges?
8. Why are frequency multipliers commonly used with crystal controlled oscillators?
9. In the frequency doubler of figure 9–2, what is the relative magnitude of the capacitance of $C2$ compared with that of $C1$?
10. How does the output of a frequency multiplier vary with the extent of frequency multiplication?
11. What three important conditions must prevail in an amplifier in order to obtain frequency multiplication?
12. In the class-C amplifier, why is energy supplied to the plate tank circuit with minimum plate losses?
13. In which element of an electron-tube amplifier is grid current flow used to develop grid-leak bias?
14. How does a loss of driving voltage endanger an electron-tube amplifier employing only grid-leak bias?
15. How do most transmitter tubes compare with receiving tubes in relative size and weight?
16. What action is prevented from occurring when an amplifier stage is properly neutralized?
17. Why are power-amplifier tetrodes and pentodes less efficient than triodes?
18. What is the term applied to certain undesirable secondary oscillations, occurring in an oscillator or amplifier, that are not the same as the fundamental or any of its harmonics?
19. How may high-frequency parasitic oscillations be reduced in a transmitter?
20. What type of keying affords complete cutoff of plate current when the key is up.

21. In figure 9–17, why is $S2$ arranged so that it can be energized only after $S1$ is closed?
22. What are the effects of not properly tuning a transmitter r-f stage?
23. What is the effect of a gassy tube on the plate and grid current of an amplifier stage?
24. Name four advantages of c-w transmission over radiotelephony.
25. What is the relative efficiency of all microphones?
26. Name two desirable qualities that a microphone should have if the frequency response is to be satisfactory?
27. Why should low-impedance microphones be used with long transmission lines?
28. What are the usual input and output zero-db reference levels of microphones?
29. What is the advantage of using a microphone of high sensitivity?
30. What are the disadvantages of using a carbon microphone?
31. Why may dynamic microphones be used with long transmission lines?
32. What are the disadvantages of a crystal microphone?
33. What desirable characteristics do magnetic microphones have?
34. In an a-m radiotelephone transmitter, how is the audio component blocked from the antenna?
35. Why is frequency multiplication necessary in the f-m transmitter shown in figure 9–25?

CHAPTER 10

TRANSMISSION LINES

INTRODUCTION

The transmission line (or antenna feedline, as it is assumed to be in this chapter) conducts or guides electrical energy from the input, or transmitter, end of the line to the output, or antenna, end of the line. If this function is to be performed with a minimum of loss, such factors as impedance matching and line losses must be considered.

Transmission lines may be classified as resonant or non-resonant lines, each of which may have advantages over the other under a given set of circumstances. There are various types of transmission lines such as the parallel two-wire line, the twisted pair, the coaxial line, and waveguides. The use of a particular type is dependent on the frequency, the voltage, the amount of power, the efficiency required, or the kind of installation to be used.

Resonant lines may have important uses other than the transmission of power. Among other uses, they may be employed as impedance-matching devices, phase shifters and inverters, wave filters and chokes, and oscillator frequency controls.

Of primary importance in the study and application of transmission lines is the characteristic impedance of the line.

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CHARACTERISTIC IMPEDANCE OF A TRANSMISSION LINE

In circuits that contain inductors and capacitors, the inductance and capacitance are present in definite "lumps." In an r-f transmission line, however, these quantities are distributed throughout the entire line and cannot be separated from each other.

The CHARACTERISTIC IMPEDANCE (or SURGE IMPEDANCE) of a transmission line having infinite length is the impedance in ohms at the operating frequency, presented by the line to the source feeding the line. This impedance across the input of a theoretically infinite line has a very valuable use. If a load equal to this impedance is connected to the output end of the line, regardless of the length of the line, the impedance presented to the source by the input terminals of the line is still equal to the characteristic impedance of the transmission line. Only one value of impedance for any particular type and size of line acts in this way.

A section of two-wire transmission line of unit length has a certain amount of resistance (no material is a perfect conductor) that varies directly with the length and inversely with the cross section of the conductor.

The same section of line has the property of distributed inductance. This property exists because of magnetic flux linkages which are established within the section when current flows. For example, an open line composed of two No. 12 conductors spaced 6 inches apart has an inductance of approximately 0.6 microhenry per foot.

This section of line also has the property of capacitance because the two wires, separated by a dielectric, act as the two plates of a capacitor. The capacitance of the two-wire line in the previous example is approximately 1.7 micromicrofarads per foot.

Finally, the transmission line of unit length has leakage resistance in the path through the insulating material that separates the two conductors (no substance is a perfect insulator). For convenience in working out problems deal-

ing with longer lines, this property usually is expressed as the reciprocal of the leakage resistance, which is conductance. The conductance is of the order of a few micromicromhos per foot.

Resistances $R1$ and $R2$, inductances $L1$ and $L2$, capacitance C, and conductance G of a unit length of two-wire transmission line are shown in figure 10–1, A. In many cases the effect of conductance G is very small compared with that produced by the inductance and capacitance and may therefore be neglected. Conductance G and resistances $R1$ and $R2$ in figure 10–1, B, are omitted and $L1$ and $L2$ are treated as if they were in one side of the transmission line.

In any circuit such as the one shown in figure 10–1, some current will flow if a voltage is applied across terminals A and B. The ratio of the voltage to the current is the impedance, Z—that is

$$Z=\frac{E}{I}.$$

The impedance presented to the input terminals of a transmission line is much more than the simple resistance of the wires in series with the impedance of the load. The effects of series inductance and shunt capacitance distributed along the line are appreciable at the relatively high operating frequency and constitute the principal components of the equivalent network.

The formula for the characteristic impedance as a function of the L and C of a unit length of transmission line may be determined from the simplified equivalent T-network circuit shown in figure 10–1, B. The conductor resistance and the insulation leakage conductance are low and considered negligible, hence are not shown in the figure. The distributed inductance of the line is divided equally in two parts in horizontal arms of the T. The distributed capacitance of the line is lumped in one value in the central leg of the T. The line is terminated in a resistive load having a value equal to that of the characteristic impedance of the line as seen looking into the T-network terminals, A and B.

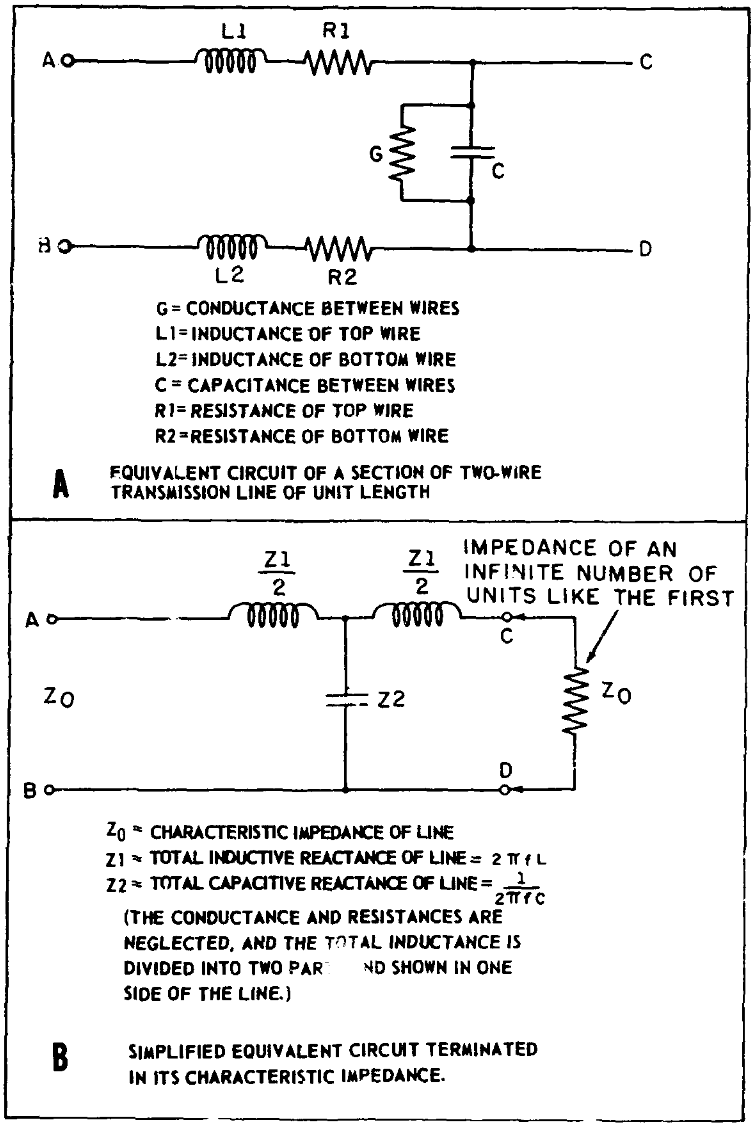


Figure 10–1.—Equivalent circuits of a two-wire transmission line of unit length.

The impedance, Z_o, looking into the T-network terminals, AB, is

$$Z_o=\frac{Z_1}{2}+\frac{Z_2\left(\frac{Z_1}{2}+Z_o\right)}{Z_2+\frac{Z_1}{2}+Z_o},$$

$$=\frac{Z_1}{2}+\frac{\frac{Z_1Z_2}{2}+Z_oZ_2}{Z_2+\frac{Z_1}{2}+Z_o},$$

$$=\frac{Z_1Z_2+\frac{Z_1^2}{2}+Z_1Z_o+\frac{2Z_1Z_2}{2}-2Z_oZ_2}{2\left(Z_2+\frac{Z_1}{2}+Z_o\right)}.$$

If both sides of this equation are multiplied by the denominator of the right-hand side, the result is

$$2Z_2Z_o+\frac{2Z_1Z_o}{2}+2Z_o^2=Z_1Z_2+\frac{Z_1^2}{2}+Z_1Z_o+\frac{2Z_1Z_2}{2}+2Z_oZ_2,$$

and this equation simplified becomes

$$2Z_o^2=2Z_1Z_2+\frac{Z_1^2}{2},$$

or

$$Z_o^2=Z_1Z_2+\left(\frac{Z_1}{2}\right)^2.$$

If the transmission line is to be accurately represented by an equivalent network, the T-network section of figure 10-1, B, must be replaced with an infinite number of similar sections. Thus, the distributed inductance of the line will be divided into n sections, instead of 2 as indicated in the last term of the preceding formula. As the number of sec-

tions approaches infinity, the last term, $\left(\frac{Z_1}{n}\right)$, will approach zero as a limit—that is, as $n \rightarrow \infty$, $\left(\frac{Z_1}{n}\right)^2 \rightarrow 0$. Therefore,

$$Z_o = \sqrt{Z_1 Z_2}\ ,$$

$$= \sqrt{\frac{2\pi fL}{2\pi fC}},$$

and

$$Z_o = \sqrt{\frac{L}{C}}.$$

The last formula indicates that the characteristic impedance depends on the distributed inductance and capacitance of the line. An increase in the separation of the wires increases the inductance and decreases the capacitance. This effect takes place because the effective inductance is proportional to the flux which may be established between the wires. If the two wires carrying current in opposite directions are placed farther apart, more magnetic flux is included between them (they cannot cancel their magnetic effects as completely as they could if the wires were closer together) and the distributed inductance is increased. The capacitance is of course lowered if the plates of the capacitor (in this case the plates are the two wires) are more widely separated.

Thus, the effect of increasing the spacing of the wires is to increase the characteristic impedance, because the $\frac{L}{C}$ ratio is increased. Similarly, a reduction in the diameter of the wires also increases the characteristic impedance. The reduction in the size of the wire affects the capacitance more than the inductance, for the effect is equivalent to decreasing the size of the plates in a capacitor in order to decrease the capacitance. Any change in the dielectric material between the two wires also changes the characteristic impedance. Thus, if a change in dielectric material increases the capacitance between the wires, the characteristic impedance is reduced.

The characteristic impedance of a two-wire line with air as the dielectric may be obtained from the formula

$$Z_o = 276 \log_{10} \frac{b}{a},$$

where b is the spacing between the centers of the conductors and a is the radius of one of the conductors.

The characteristic impedance of a concentric, or coaxial, line also varies with L and C. However, because the difference in construction of the two lines causes L and C to vary in a slightly different manner, the following formula must be used in determining the characteristic impedance of the concentric line:

$$Z_o = 138 \log_{10} \frac{b}{a},$$

where b is the inner diameter of the outer conductor and a is the diameter (or the outer diameter, if a hollow tube is used) of the inner conductor.

WAVE MOTION ON AN INFINITE LINE

Figure 10–2 shows sine waves of voltage and current that travel at high speed along a two-wire transmission line of infinite length. Because this is a line of infinite length, no reflections occur; therefore, the voltage and the current are in phase with each other everywhere along the line. Because of line losses, the curves diminish in amplitude as the waves progress along the line. If a voltage is impressed on a line such as this, an electric field will be established between the wires. Likewise, current will flow in the wires, and a magnetic field will be established around each wire. These two fields constitute an electromagnetic wave that travels down the wire at a velocity somewhat less than that of light.

Figure 10–2 illustrates what would happen if the voltage and current could be stopped for an instant in time. An instant later all waves would have moved to the right a

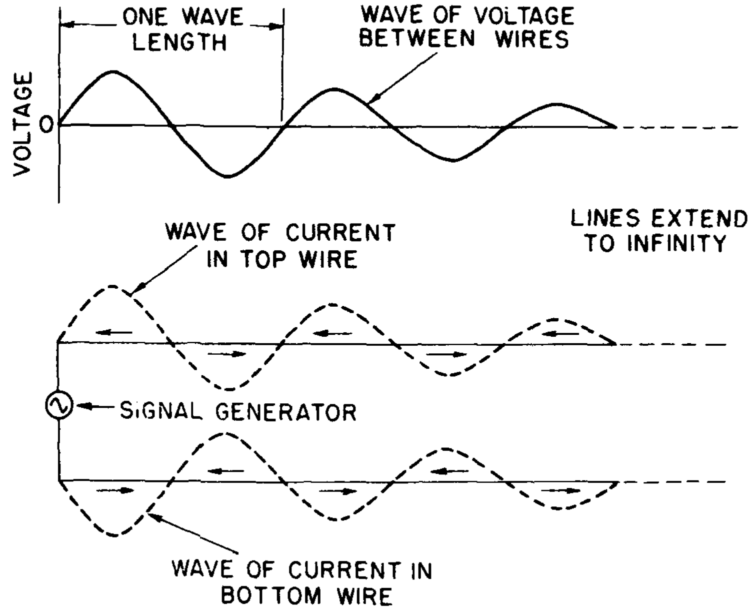


Figure 10–2.—Traveling waves of current and voltage on a line of infinite length.

slight amount. In this figure the waves are stopped at the instant when the alternating source voltage has just reached zero.

Traveling waves exist on the line because it takes a certain amount of time to propagate them down the line. The production of these waves may be better understood from the following considerations. First of all, it must be understood that the waveforms shown in figure 10–2 are "stopped" for an instant in time and the observer examines the entire train of waves along the line. Assume that at a given instant the voltage at the generator terminals is zero. An instant later one terminal becomes more positive and the other becomes more negative. The electric field between the wires increases in strength; the current and also the magnetic field increase proportionately. The perpendicular

distance from any point along the wire to the current curve (fig. 10–2) indicates the relative magnitude and direction of the current at that point. The perpendicular distance from any point along the voltage axis to the voltage curve represents the relative magnitude and polarity of the voltage across the line at the corresponding location.

At 90° in the electrical cycle the electric and magnetic field are at their maximum, and from 90° to 180° they decrease in amplitude to zero. At 180° the voltage at the generator terminals reverses polarity, and the electric field between the wires reverses direction. Similarly, the current reverses direction, and this causes the magnetic field to reverse direction. The fields increase in strength from 180° to 270°, and then decrease in strength to zero at 360°. These electric and magnetic impulses do not return to the generator once they start down a line of infinite length.

The characteristics of a theoretically infinite line may be summarized as follows:

1. The voltage and current are in phase throughout the line.
2. The ratio of the voltage to the current is constant over the entire line and is known as the CHARACTERISTIC IMPEDANCE.
3. The input impedance is equal to the characteristic impedance.
4. Since the voltage and current are in phase, the line operates at maximum efficiency.
5. Any length of line can be made to appear like an infinite line if it is terminated in its characteristic impedance.

LINE REFLECTIONS

If a transmission line is infinitely long, or if it is terminated in its characteristic impedance, reflections do not occur. However, if there is an abrupt discontinuity in the line (such as an open circuit or a short circuit) complete reflection will occur. A discontinuity of less importance (such as a poorly made splice) will cause some reflection, the amount of re-

flection depending on the value of the resistance at the splice. In this section, the discussion of reflection will be limited to the two extreme conditions—that is, to open- and closed-end lines.

Open-End Lines

One type of r-f transmission line is the open-end line, in which the impedance at the output end can be considered as practically infinite because no load is attached. When energy is applied to the generator end, the first surge consists of a wave of current and a wave of voltage that sweep down the line in phase with each other—that is, their positive maxima are together. The initial current and voltage waves must travel down the line in phase because the characteristics of the line are the same as those of a line that is truly infinite WHILE THE INITIAL WAVE IS TRAVELING TOWARD THE OUTPUT END. The in-phase condition of these waves can be changed only when they encounter a difference in the impedance between the two wires of the line, and reflections occur. Thus, when the wave of current reaches the open-circuited output end (terminal point) of the line the current must collapse to zero. When the current wave collapses, the magnetic field that was set up by it also collapses. The collapsing magnetic field cuts the conductors near the output end and induces additional voltage across the line. This voltage acts, in a way, like a reverse generator and sets up new current and voltage waves that travel back along the line toward the input end.

An open-end transmission line one wavelength long and having no attenuation is shown in figure 10–3, A. It should be pointed out that all of the waveforms shown are instantaneous values that exist along a transmission line. These curves are unlike conventional sine curves in which distance along the X axis represents lapse in time from the origin, proceeding to the right. Instead, these curves represent instantaneous values of current or voltage as they exist all along the line for the same instant of time.

Although only four positions of the generator voltage

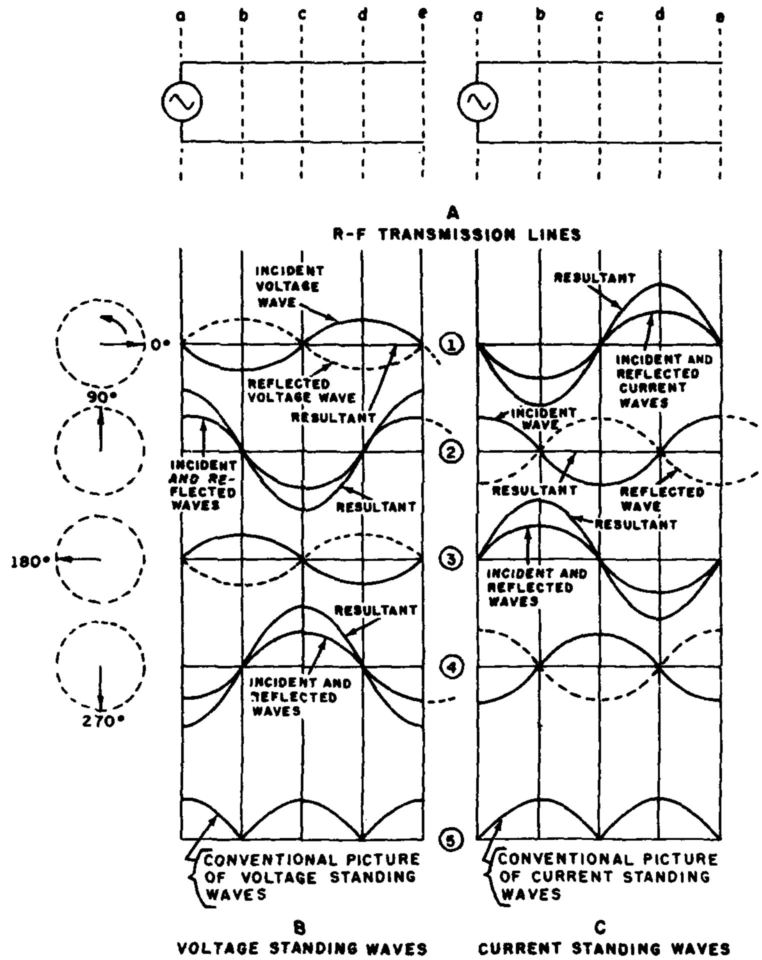


Figure 10–3.—Formation of standing waves on an open-end transmission line one wavelength long.

vector are shown, the picture could be made more complete by showing the waveforms at 45° intervals. At intervals of 45° between the generator positions shown—for example, at 45°, 135°, 225°, and 315°—the instantaneous values of the resultants will be 0.707 of their maximum values. The line is represented twice (fig. 10–3, A) in order to orient it properly with respect to the waves of voltage and current. These waves are shown separately (fig. 10–3, B and C) in order to simplify the analysis, although in reality they both appear simultaneously along the same transmission line.

VOLTAGE STANDING WAVES.—It is assumed that in part ① of figure 10–3, B, the generator voltage vector has gone through at least two complete revolutions so that the voltage wave has had time to travel down the line and return to the generator end. The waveforms are stopped in time in this figure at the instant that the generator voltage vector is at the zero position.

It may be observed that the initial voltage wave is reflected at the output end of the line in phase with the voltage wave that would have continued along the line in the original direction of travel if the line had been longer. For example, the dotted waveform extending slightly beyond the end of the line in part ① indicates that the incident wave would have started going negative. Therefore, the reflected wave will start back in a negative direction. At the instant in time being considered here the incident and reflected wave add vectorially to give a zero resultant wave.

Ninety degrees later (part ② of fig. 10–3, B) the incident and reflected waves are in phase and add vectorially to give the resultant voltage wave, as shown. At 180° (part ③), the resultant voltage is again zero; and at 270° (part ④), the incident and reflected waves are once again in phase, and the resultant voltage wave is shown 180° out of phase with the resultant wave in part ②.

Next, consider the voltage variations across the line that occur with respect to time at certain locations along the line. At point *a*, the voltage is zero (part ①), then maximum in one direction (part ②), then zero again (part

③), and finally maximum in the other direction (part ④). This is true also at points *c* and *e*; at points *b* and *d* the voltage is always zero. A suitable voltage-indicating device (to be discussed later in the chapter) located at *a*, *c*, or *e* will indicate voltage loops (points of maximum voltage); and at *b* and *d* the device will indicate voltage nodes (points of minimum voltage).

Standing waves of voltage are shown in part ⑤ of figure 10–3, B. This curve represents effective values of voltage at the various points along the line. These values are actually the effective values of the sinusoidal voltage variations occurring across the line at the points where the measurements are being made. Thus, at point *c* the voltage will be zero at one instant of time (part ①), then it will build up to a maximum with one polarity (part ②), then it will become zero (part ③), and finally it will build up to a maximum with the opposite polarity (part ④).

Current standing waves.—Standing waves of current are shown in figure 10–3, C. They are occurring simultaneously with the voltage waves on the transmission line, but they are shown here separatelv in order to simplify the figure. The initial current wave is reflected at the output end of the line 180° out of phase with the current wave that would have continued along the line in the original direction of travel if the line had been extended. (See dotted-line extensions.) In other words, the reflected current reverses direction at the open end of the line and is shown 180° out of phase with the incident wave except when the incident and reflected waves have zero values at the end of the line (0°, 180°, and so forth). This reversal is opposite to the condition for voltage, because the reflected voltage wave has the same polarity that the incident wave would have if it continued down the extended line in the direction of travel. (See dotted-line extensions.)

Because the incident and reflected current waves are 180° out of phase at the open end of the line, they cancel at this point and the resultant current at the open end is always zero. The rotating vectors at the left of the figures indicate

the generation of sine waves of both voltage and current. At *b* the incident and reflected current waveforms combine to produce a current loop (maximum current); the same is also true at point *d*. At *a*, *c*, and *e* the incident and reflected waveforms combine to produce current nodes (points of zero current).

In part ② of figure 10–3, C, the current vector has rotated 90°. Combining the incident and reflected waves at this instant gives a resultant current waveform that has an amplitude of zero throughout the entire length of the line. In part ③ of the figure the current vector has completed 180° of its cycle. Again combining the incident and reflected waves at this instant gives a resultant current that has a maximum amplitude at points *b* and *d* although the direction of the current is opposite to that in part ① at these points. In part ④ the resultant current is again zero all along the line.

Part ⑤ of figure 10–3, C, is a plot of the EFFECTIVE values of current at the various points along the line. These current values are actually the effective values of the resultant current variations through the cycle at the respective points where the measurements are made. For example, at point *b* the current is maximum in one direction at one instant of time (part ①), then it becomes zero (part ②), then it builds up to a maximum in the opposite direction (part ③), and finally returns to zero (part ④). The effective value of this current variation is plotted at *b* in part ⑤.

Closed-End Lines

In a closed-end line the voltage and current waveforms exchange places with respect to their locations in an open-end line. Thus, in figure 10–3, A, if the line is closed at the end, part B becomes current standing waves and part C becomes voltage standing waves. At the short-circuited end of the line (point *e*), the current varies from zero to a maximum in one direction and back to zero and a maximum in the other direction. The effective value of the current will therefore

be a maximum at point *e* (a current loop). The voltage ($E=IR$) across the short circuit approaches zero because the resistance of the short circuit is negligible.

At $\frac{\lambda}{4}$ and $\frac{3\lambda}{4}$ wavelengths from the shorted end, the effective current is a minimum at all times and the effective voltage is a maximum. At $\frac{\lambda}{2}$ and λ wavelengths from the shorted end, the effective value of the current is a maximum and the effective voltage is zero. The curves representing the variation of effective current and voltage along a shorted line are shown in figure 10–4.

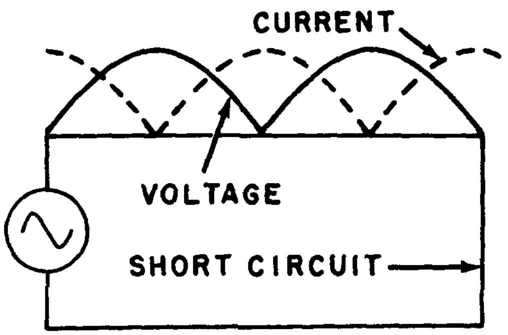


Figure 10–4.—Conventional picture of current and voltage standing waves on a closed-end line.

NONRESONANT LINES

A nonresonant line can be defined as a line that has no standing waves of current and voltage. Such a line is either infinitely long or is terminated in its characteristic impedance. Because there are no reflections, all of the energy passed along the line is absorbed by the load. The voltage and current waves are traveling waves that move in phase with each other from the source to the load.

On lines carrying radio frequencies, the characteristic impedance is almost always pure resistance. Therefore, it is customary to say that a nonresonant line is terminated in

a resistive load equal to the characteristic impedance of the nonresonant line.

RESONANT LINES

A resonant transmission line is one that has standing waves of current and voltage. The line is of finite length and is NOT terminated in its characteristic impedance, and therefore reflections are present.

A resonant line, like a tuned circuit, is resonant at some particular frequency. The resonant line will present to its source of energy a high or a low resistive impedance at multiples of a quarter-wavelength. Whether the impedance is high or low at these points depends on whether the line is short- or open-circuited at the output end. At points that are not exact multiples of a quarter-wavelength, the line acts as a capacitor or an inductor.

A resonant transmission line thus may assume many of the characteristics of a resonant circuit that is composed of lumped inductance and capacitance. The more important circuit effects that resonant transmission lines have in common with resonant circuits having lumped inductance and capacitance are as follows:

1. SERIES RESONANCE—Resonant rise of voltage across the reactive circuit elements, and low impedance across the resonant circuit.
2. PARALLEL RESONANCE—Resonant rise of current in the reactive circuit elements, and high impedance across the resonant circuit.

Resonance in Open-End Lines

The open-end resonant line may be better understood by means of an analysis of figure 10–5. The transmission lines considered in this figure and in the next have no losses. If losses are present (and they are in a practical line), the voltage at voltage nodes is not zero; neither is the current zero at current nodes. However, in these figures losses are neglected in order to simplify the analysis. Figure 10–5

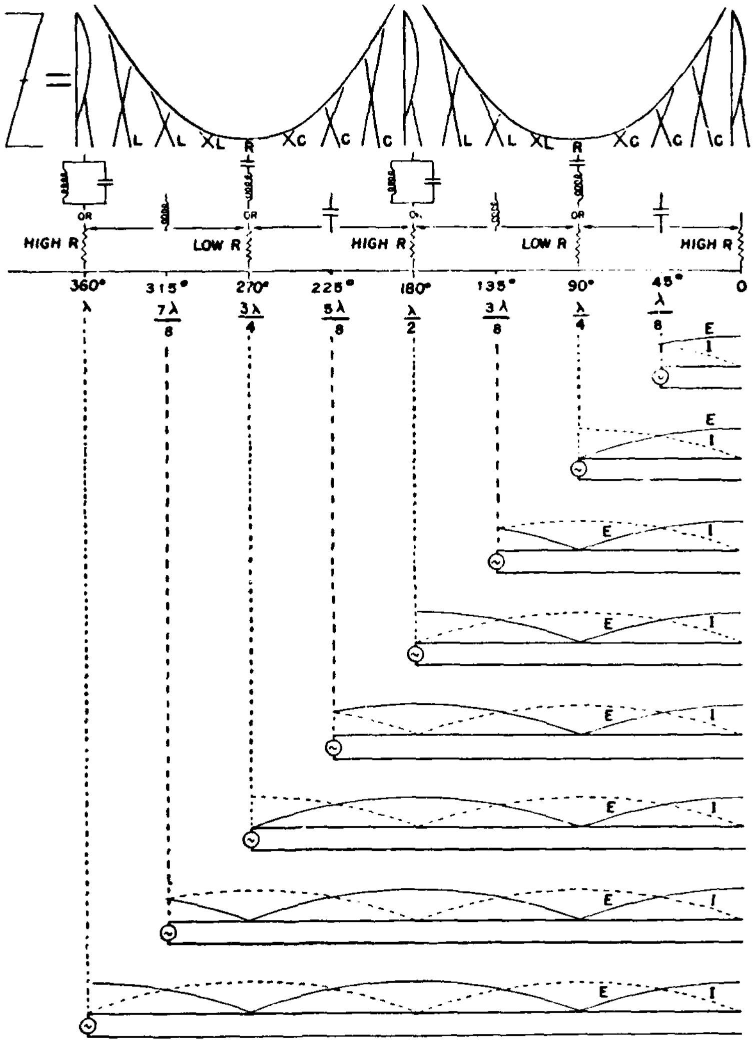


Figure 10-5.—Impedance characteristics of open-end resonant lines.

illustrates the relation of voltage, current, and impedance for various lengths of open-end transmission lines. The impedance that the generator "sees" at various distances from the output end is shown directly above in the impedance curves. The curves above the letters (R, X_L, X_C) of various heights indicate the relative magnitudes of the impedances presented to the generator for the various lengths of lines indicated. The letters themselves indicate the type of impedance offered at the corresponding inputs. The circuit symbols above the various transmission lines indicate the equivalent electrical circuits for the transmission line at that particular length (measured from the output end). The curves of effective E and I whose ratio, $\frac{E}{I}$, is the impedance, Z, are shown above each line.

At all ODD quarter-wavelength points $\left(\frac{\lambda}{4}, \frac{3\lambda}{4}, \frac{5\lambda}{4}, \text{etc.}\right)$ measured from the OPEN END of the line, the current is a maximum and the impedance is a minimum. In addition, there is a resonant rise of voltage from the odd quarter-wavelength points toward the open end. Thus at all odd quarter-wavelength points the open-end transmission line acts like a series resonant circuit. The impendance is therefore very low and is prevented from being zero only by the small circuit losses.

At all EVEN quarter-wavelength points $\left(\frac{\lambda}{2}, \lambda, \frac{3\lambda}{2}, \text{etc.}\right)$ the voltage is a maximum, and therefore the impedance is a maximum. A comparison of this type of transmission line with an L-C resonant circuit shows that at even quarter-wavelengths (from the output end) the line ACTS like a parallel resonant circuit.

In addition to acting as series or parallel L-C resonant circuits, resonant open-end lines also may act as nearly pure capacitances or inductances when the lengths of the lines are not an exact multiple of the fundamental quarter-wavelength corresponding to the frequency of the applied voltage at the input terminals. Figure 10-5 shows that an

open-end line less than a quarter-wavelength long acts like a capacitance; between $\frac{\lambda}{4}$ and $\frac{\lambda}{2}$ wavelength, as an inductance; between $\frac{\lambda}{2}$ and $\frac{3\lambda}{4}$ wavelength, as a capacitance; between 4 and λ wavelength, as an inductance; and so forth.

Resonance in Closed-End Lines

The closed-end line may likewise be studied with the aid of figure 10–6. At ODD quarter-wavelengths from the closed end of the line the voltage is high, the current low, and the impedance high. Because conditions are similar to those in a parallel resonant circuit, the shorted transmission line of odd quarter-wavelengths acts like a parallel resonant circuit. The voltage across a circuit of this type cannot exceed the applied voltage.

At EVEN quarter-wavelength points (measured from the shorted end) the voltage is a minimum, the current is a maximum, and the impedance is a minimum. Because this action is similar to series resonance in an L-C circuit, a shorted transmission line of even quarter-wavelengths acts like a series resonant circuit.

Resonant closed-end lines, like open-end lines may also act as nearly pure capacitances or inductances when the length of the lines are not exact multiples of the fundamental quarter-wavelength corresponding to the frequency of the applied voltage at the input terminals.

Line Terminated in a Reactance

A line terminated in a RESISTANCE equal to its characteristic impedance normally has no reflections present. However, if a transmission line is terminated in a REACTANCE equal to its characteristic impedance or to any other impedance, standing waves are NOT eliminated. Figure 10–7, A, shows the standing waves that exist on a line terminated in a capacitive reactance equal to its characteristic impedance. Note that the last current loop is less than a quarter-wavelength from the capacitive termination of

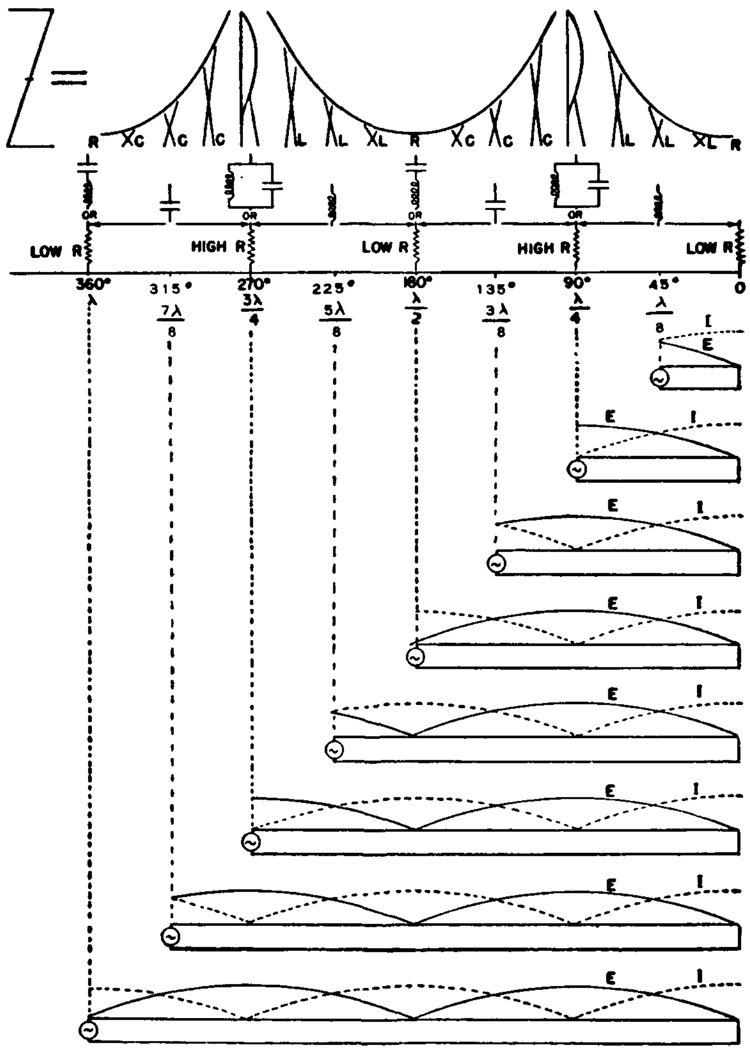


Figure 10-6.—Impedance characteristics of closed-end resonant lines.

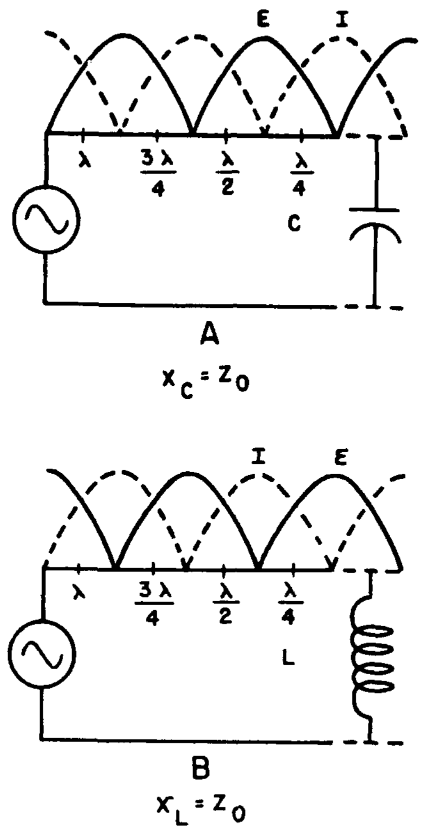


Figure 10–7.—Transmission lines terminated in reactances.

the line. With capacitive termination the voltage and current distribution has essentially the same character as with the open-end line, except that the curves are shifted toward the output end of the line by an amount that increases as the capacitive reactance is reduced—that is, as the line approaches the closed-end condition of zero impedance.

Figure 10–7, B, shows the standing waves that occur on a line terminated in an inductive reactance equal to the

characteristic impedance. Note that the last voltage loop is less than a quarter-wavelength from the inductive termination of the line. With inductive termination the voltage and current distribution has essentially the same character as with a short-circuited output, except that the curves are shifted toward the inductive termination by an amount that increases as the terminating inductive reactance approaches infinity—that is, as the line approaches the open-end condition.

Standing-Wave Ratio

The ratio of the effective voltage at a loop to the effective voltage at a node, or the effective current at a loop to the effective current at a node is called the STANDING-WAVE RATIO (SWR) of a transmission line. It is also equal to the ratio of the characteristic impedance of the line to the impedance of the load, or vice versa. When the line is terminated in a perfect match, all of the energy sent down the line is abosrbed by the load and none is reflected. Under these conditions no standing waves are present. The maximum and minimum values are the same, and therefore the standing-wave ratio is equal to 1.0.

Two mismatched lines are shown in figure 10–8. In each of these lines the characteristic impedance, Z_o, of the line is 300 ohms. In figure 10–8, A, the load impedance is 60 ohms. The ratio of the effective current at a to the effective current at b is equal to $\frac{y}{x}$, or $\frac{5}{1}$, which is also equal to $\frac{300}{60}$.

In figure 10–8, B, the line impedance is less than $\left(\frac{1}{5}\text{ of}\right)$ the impedance of the load. The ratio of the effective current at d to the effective current at e is equal to $\frac{1,500}{300}$, or $\frac{5}{1}$. In the first example, the SWR is equal to the ratio of the Z_o of the line to the Z of the load. In the second example it is equal to the ratio of the Z of the load to the Z_o of the line. In both examples, the SWR is equal to the ratio of the effective current at a loop to the effective current at a node.

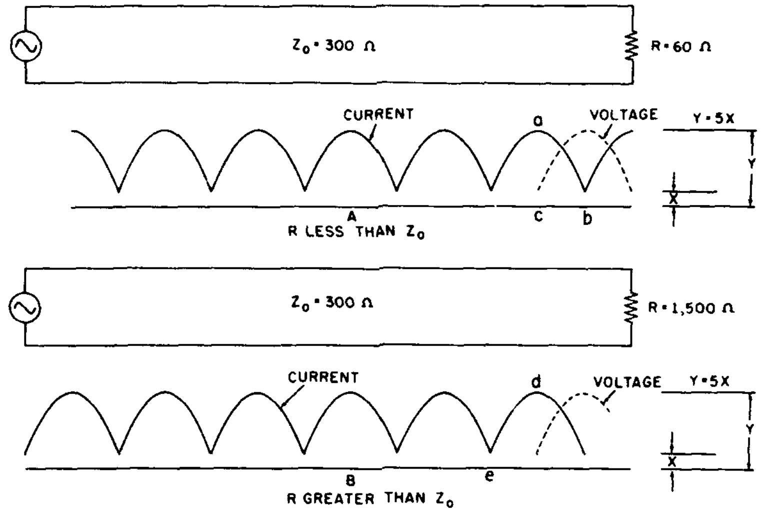


Figure 10–8.—Mismatched lines showing standing-wave ratio.

In general, the higher the SWR, the greater is the mismatch between the line and the load. A knowledge of the position of the current and voltage loops and nodes along the line will indicate whether the load resistance is less than or greater than the characteristic impedance. For example, in figure 10–8, A, there are a voltage node and a current loop at the load. This occurs because the load resistance is less (approaching a shorted condition) than the characteristic impedance of the line. Thus, it is a simple matter (by the use of one of the r-f measuring devices to be discussed later) to determine whether the load resistance is greater or smaller than Z_o. If the load resistance is greater than Z_o (fig. 10–8, B), the output end of the line will appear more like an open circuit, and r-f measuring devices will indicate maximum effective voltage and minimum effective current at that point.

TYPES OF TRANSMISSION LINES

There are five general types of transmission lines—the parallel two-wire line, the twisted pair, the shielded pair, the concentric (coaxial) line, and waveguides. As mentioned in the introduction, the use of a particular type of line depends among other things on the frequency and the power to be transmitted, and on the type of installation.

Parallel Two-Wire Line

One of the most common types of transmission lines consists of two parallel conductors that are maintained at a fixed distance by means of insulating spacers or spreaders that are placed at suitable intervals. This type of line is shown in figure 10–9, A. The line is used frequently because of its ease of construction, its economy, and its efficiency. In practical applications two-wire transmission lines (with individual insulators rather than spacers) are used for power lines, rural telephone lines, and telegraph lines. This type of transmission line is also used as the connecting link between an antenna and transmitter or an antenna and receiver.

In practice, such lines used in radio work are generally spaced from 2 to 6 inches apart on 14-mc and lower frequencies. The maximum spacing for 38-mc or higher frequencies is 4 inches. In any case, in order to effect the best cancellation or radiation, it is necessary that the wires be separated by only a small fraction of a wavelength. For best results, the separation should be less than 0.01 wavelength.

The principal disadvantage of the parallel-wire transmission line is that it has relatively high radiation loss and therefore cannot be used in the vicinity of metallic objects, especially when high frequencies are used, because of the greatly increased loss which results.

Uniform spacing of a two-wire transmission line may be assured if the wires are imbedded in a solid low-loss dielectric throughout the length of the line, as indicated in figure 10–9, B. This type of line is often called a TWO-WIRE

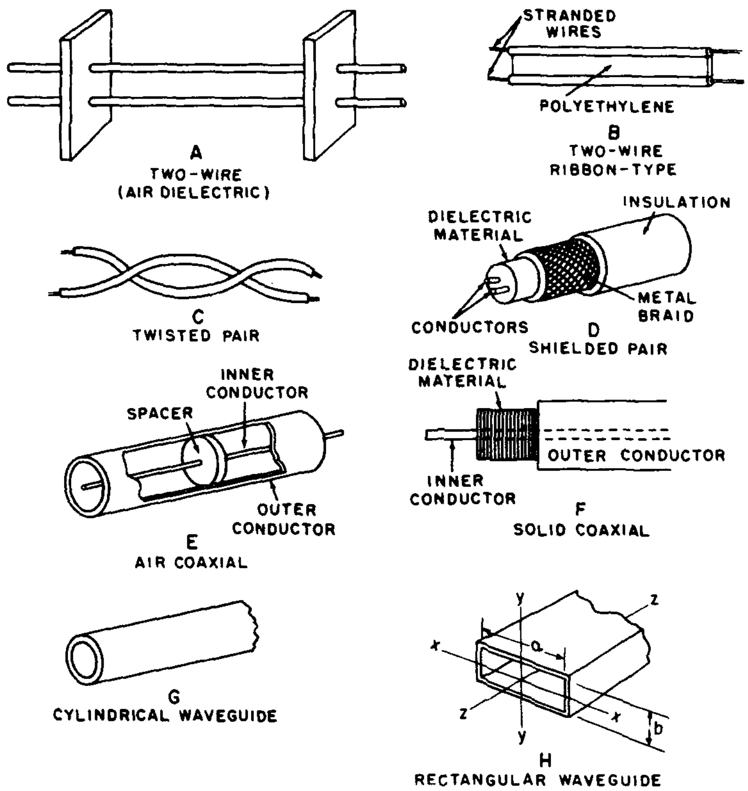


Figure 10-9.—Types of transmission lines.

RIBBON type. The ribbon type is commonly made with two characteristic impedance values, 300 ohms and 75 ohms. The 300-ohm line is about one-half inch wide and is made of stranded wire. Because the wires are imbedded in only a thin ribbon of polyethylene, the dielectric is partly air and partly polyethylene. Moisture or dirt will change the characteristic impedance of the line. This effect becomes more serious if the line is not terminated in its characteristic impedance.

The wires of the 75-ohm line are closer together, and the field between the wires is confined largely to the dielectric. Weather and dirt therefore affect this line less than they affect the 300-ohm line. The ribbon-type of line is widely used to connect television receivers to their antennas.

Twisted Pair

The twisted pair is shown in figure 10–9, C. As the name implies, it consists of two insulated wires twisted to form a flexible line without the use of spacers. It is used as an untuned line (on a tuned line the insulation might be punctured at voltage loops) for low-frequency transmission. It is not used for the higher frequencies because of the high losses occurring in the rubber insulation. When the line is wet, the losses increase greatly. The characteristic impedance of such lines is about 100 ohms, depending on the type of cord used.

Shielded Pair

The shielded pair (shown in fig. 10–9, D) consists of two parallel conductors separated from each other and surrounded by a solid dielectric. The conductors are contained within a copper-braid tubing that acts as a shield. This assembly is covered with a rubber or flexible composition coating to protect the line against moisture and friction. Outwardly, it looks much like an ordinary power cord for an electric motor.

The principal advantage of the shielded pair is that the two conductors are balanced to ground—that is, the capacitance between each conductor and ground is uniform along the entire length of the line and the wires are shielded against pick-up of stray fields. This balance is effected by the grounded shield that surrounds the conductors at a uniform spacing throughout their length.

If radiation from an unshielded line is to be prevented, the current flow in each conductor must be equal in amplitude in order to set up equal and opposite magnetic fields that are

thereby canceled out. This condition may be obtained only if the line is clear of all obstructions, and the distance between the wires is small. If, however, the line runs near some grounded or conducting surface, one of the two conductors will be nearer that obstruction than the other. A certain amount of capacitance exists between each of the two conductors and the conducting surface over the length of the line, depending upon the size of the obstruction. This capacitance acts as a parallel conducting path for each half of the line, causing a division of current flow between each conductor. Since one conductor may be nearer the obstruction than the other, the current flow will accordingly be increased, resulting in an inequality of current flow in the two conductors and therefore incomplete cancellation of radiation. The shielded line, therefore, eliminates such losses to a considerable degree by maintaining balanced capacitances to ground.

Air Coaxial

The air coaxial line has advantages that make it practical for operation at the ultrahigh frequencies. It consists of a wire mounted inside of, and coaxially with, a tubular outer conductor (fig. 10–9, E). In some cases the inner conductor also is tubular. The inner conductor is insulated from the outer conductor by insulating spacers or beads at regular intervals. The spacers are made of pyrex, polysterene, or some other material possessing good insulating qualities and having low loss at high frequencies.

The chief advantage of the coaxial line is its ability to keep down radiation losses. In the two-wire parallel line the electric and magnetic fields extend into space for relatively great distances and tend to cause radiation losses and noise pick-up from other lines. In a coaxial line, however, no electric or magnetic fields extend outside the outer conductor. They are confined to the space between the two conductors. Thus, the coaxial line is a perfectly shielded line.

The disadvantages of such a line are: (1) It is expensive;

(2) at extremely high frequencies, its practical length is limited because of the considerable loss that occurs; and (3) it must be kept dry in order to prevent excessive leakage between the conductors. To prevent condensation of moisture, the line may be filled in certain applications with dry nitrogen at pressures ranging from 3 to 35 pounds per square inch. The nitrogen is used to dry the line when it is first installed, and a pressure is maintained to ensure that the leakage will be outward.

Solid Coaxial

Concentric cables are also made with the inner conductor consisting of flexible wire insulated from the outer conductor by a solid and continuous insulating material, as shown in figure 10–9, F. Flexibility may be gained if the outer conductor is made of a metal braid, but the losses in this type of line are relatively high.

Waveguides

Two common types of waveguides are the cylindrical type (fig. 10–9, G) and the more often used rectangular type (fig. 10–9, H). The term "waveguide" is applicable to all types of transmission lines in the sense that they are used to direct or guide the energy from one point to another. In this sense it does not matter whether the line is composed of a single conductor, two or more conductors, a coaxial line, a hollow metal tube, or a dielectric rod. Usage, however, has limited the meaning of the word to the hollow metal tube and the dielectric transmission line. The term "waveguide," as used in this text, means a hollow metal tube.

The transmission of an electromagnetic wave along a waveguide is closely related to its transmission through space. At power-line frequencies the current flow through the conductors was long considered to be the means by which energy is transmitted over a line, and the external electric and magnetic fields were regarded as coincidental to that

transmission. That this may not necessarily be the case may be deduced from the fact that today energy may be transmitted along a waveguide with no longitudinal current flow along the guide. Thus it is believed that the energy transmitted is contained in the electromagnetic fields that travel down the waveguide and current flow in the guide walls only provides a boundary for these electric and magnetic fields.

TYPES OF WAVEGUIDES.—Waveguides may be classed according to cross section (rectangular, elliptical, or circular) or according to material (metallic or dielectric). Dielectric waveguides are seldom used because the losses for all known solid dielectric materials are too great for efficient transmission.

Of the three types of hollow-tube waveguides the rectangular cross-section type is most commonly used. Circular waveguides are seldom used because it is difficult to control the plane of polarization and the mode of operation. (Modes are described later.) Circular waveguides involve the further difficulty of joining curved surfaces when a junction is required. They do find use in rotating joints, however, because of their circular symmetry, both physical and electrical. Elliptical waveguides are not used because of fabrication, joining, and bending difficulties.

ADVANTAGES OF HOLLOW WAVEGUIDES.—A hollow waveguide has lower loss than either an open-wire line or a coaxial line in the frequency range for which it is practical. An open-wire line has three kinds of loss—(1) radiation loss, (2) dielectric loss, and (3) copper loss. In the coaxial line there is no radiation loss because the outer conductor acts as a shield which confines the magnetic and electric fields to the space between the inner and outer conductors. Both the coaxial line and the hollow pipe are perfectly shielded lines and therefore have no radiation loss.

Dielectric loss in the insulating beads of a coaxial line is considerable at very high frequencies, but air has negligible dielectric loss at any frequency. Because hollow metal wave-

guides are usually filled with air, they have negligible dielectric loss.

The third kind of loss is the copper loss. At high frequencies the current flows in a thin layer near the surface of the conductor. As the frequency increases, the thickness of this layer decreases, thus reducing the effective cross section of the conductor and causing the copper loss to increase as the effective resistance of the conductor becomes greater. In a coaxial line most of the resistance and most of the copper loss are in the inner conductor because the circumference of this conductor is less, and for a given penetration of current the effective cross section is less than that of the outer conductor.

For example, if the current flows in a very thin layer at the surface of both conductors, and if the inner circumference of the outer conductor is five times that of the outer surface of the inner conductor, the area through which current flows in the outer conductor is five times that of the inner conductor. The resistance, R, of a conductor is

$$R=\frac{\rho L}{A},$$

where ρ is the resistivity of the metal, L the length of the conductor, and A the area of the cross section through which the current flows. Therefore, the resistance of the inner conductor is five times that of the outer conductor. If the inner conductor were eliminated, the copper losses would be greatly reduced. A coaxial line without the inner conductor is equivalent to a round hollow waveguide.

Because the waveguide has less copper loss than a coaxial line, and because it has negligible dielectric loss and no radiation loss, the total losses of a waveguide above the cutoff frequency are less than those of a coaxial line of the same size operating at the same frequency.

The waveguide is simpler in construction than the coaxial line because the inner conductor and its supports are eliminated. Because there is no inner conductor which may be

displaced or broken by vibration or shock, the waveguide is more rugged than the coaxial line.

DISADVANTAGES OF HOLLOW WAVEGUIDES.—The minimum size of the waveguide that can be used to transmit a certain frequency is proportional to the wavelength at that frequency. This proportionality depends upon the shape of the waveguide and the manner in which the electromagnetic fields are set up within the pipe. In all cases there is a minimum frequency that can be transmitted. The lowest cutoff frequency is determined by the inside dimensions shown in figure 10-9, H. The wavelength, λ_{co}, corresponding to the cutoff frequency is equal to twice the inside width of the guide, or

$$\lambda_{co}=2a.$$

Higher frequencies, however, can be transmitted. The width of the guide for these frequencies is greater than their corresponding free-space half-wavelengths.

The distance, b, is not critical with regard to frequency. However, this distance determines the voltage level at which the waveguide arcs over. Therefore, for high power and voltage, the distance, b, should be large. In practice, b may be from 0.2 to 0.5 of the wavelength in air, and a may be about 0.7 times the wavelength in air.

Because the cutoff frequency corresponds to a wavelength that is equal to twice the inside width of the guide, waveguides are not used extensively at frequencies below approximately 3,000 mc (10 cm). At lower frequencies the guide would be too large. For example, to transmit 10-centimeter waves, a rectangular waveguide would have to be wider than 5 centimeters. For 1-meter waves the waveguide would have to be about 2⅓ feet wide, and for 10-meter waves, 23 feet wide.

The installation of a waveguide transmission system is somewhat more difficult than the installation of other types of lines. The radius of bends in the guide must be greater than two wavelengths to avoid excessive attenuation and the cross section of the guide must be maintained uniform around

the bend. These difficulties hamper installations in restricted spaces. If the guide is dented, or if solder is permitted to run inside the joints, the attenuation of the line is greatly increased. In addition to the increased attenuation that they cause, dents and beads of solder also reduce the breakdown voltage of the waveguide and cause standing waves in the guide.

Although such faults may not cause arc-over in the guide, they limit the power-handling capacity of the system and make the possibility of arc-over more likely. Thus, unless great care is exercised in the installation, one or two carelessly made joints may nullify completely the initial advantage obtained from the use of the waveguide.

MODES OF TRANSMISSION.—For convenience of reference, a system employing letters and subscript numbers was devised for describing waveguide modes. Figure 10–10, A, shows the *TE* mode of operation of a rectangular waveguide. The letters *TE* indicate a mode of operation in which the ELECTRIC field (composed of parallel *E* lines) lies in transverse planes that contain the *X* and *Y* axes and in which the *E* lines are parallel to the *Y* axis and are perpendicular to the longitudinal (*Z*) axis of the guide. Similarly the letters *TM* (fig. 10–10, B) indicate that the MAGNETIC field (composed of closed loops) lies in transverse planes that contain the *X* and *Y* axes and are wholly transverse to the guide axis.

For rectangular waveguides the accepted system of subscripts is that the first number subscript indicates the number of half-wave variations of the transverse field in the wide dimension of the guide, and the second number indicates the number of half-wave variations of the same field in the narrow dimension. The TE_{10} mode, for example, means that the electric field has one half-wave variation in the wide dimension, and none in the narrow dimension. The TM_{11} mode means that the magnetic field has one half-wave variation in both the wide and the narrow dimensions. The mode having the lowest cutoff frequency for a given size of guide is called the DOMINANT MODE for that guide. The dominant mode, TE_{10}, for rectangular waveguides is

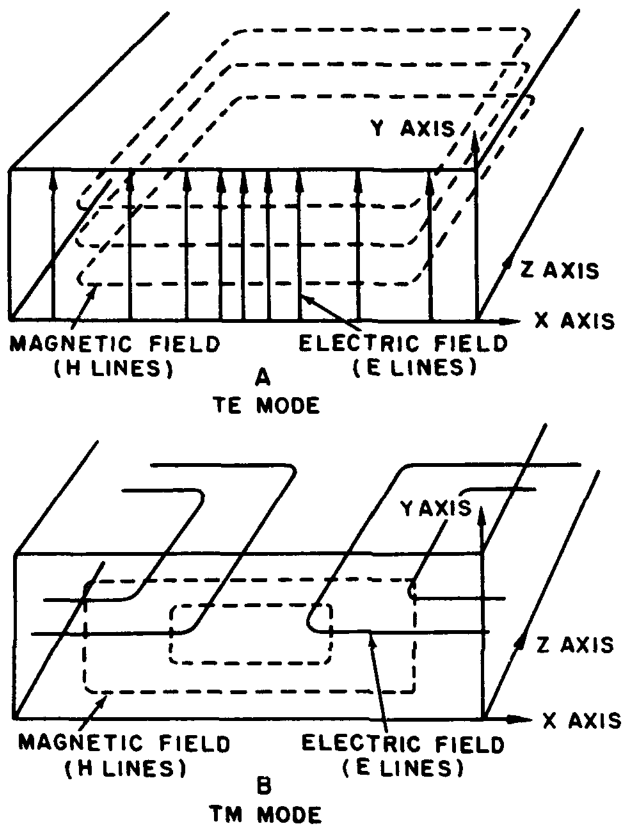


Figure 10–10.—*TE* and *TM* modes of operating a waveguide.

the mode most commonly used. There are many reasons for this. This mode is easily excited, it is plane-polarized, it is easily matched to a radiator, and the plane of polarization is easily controlled. Other reasons are that its cutoff frequency is dependent upon only one of the guide dimensions while many of the other modes depend on two dimensions, with the result that it is easy to design the waveguide so that only this one mode may exist in it.

COUPLING.—There are three principal ways in which energy can be put into and removed from waveguides. The first is by placing a small loop of wire so that it "cuts" or couples the *H* lines of the magnetic field, as in a simple transformer. The second is by providing an "antenna" or probe which can be placed parallel to the *E* lines of the electric field. The third method is to link or contact the fields inside the guide by external fields through the use of slots or holes in the walls.

The foregoing is a general description of waveguides. A fuller discussion of the theory and operation of waveguides and cavity resonators will be included in an advance course.

MEASUREMENTS ON R-F LINES

Methods of Making Measurements

It is often necessary to determine if standing waves are present on a transmission line and, if present, where the loops and nodes of voltage and current occur. It may also be necessary to determine the SWR for voltage and current. Therefore, it is necessary to make electrical measurements on the line.

There are several methods of determining the magnitude of voltage or current at any point on an r-f line. In making these measurements it must be remembered that the magnetic field around a line varies directly with the current, and that the electrostatic field about the line varies directly with the voltage.

Current may be observed at any point on a line either by cutting the line and inserting a suitable ammeter (fig. 10–11, A), or by placing a loop (fig. 10–11, B) connected to an ammeter in the magnetic field. In figure 10–11, A, the a-c ammeter is connected in series with the line. The ammeter indicates the standing wave of current at the point where the measurement is made. If this method is not practical, the method shown in figure 10–11, B, may be used instead. The coil is moved along the line, and the meter will indicate

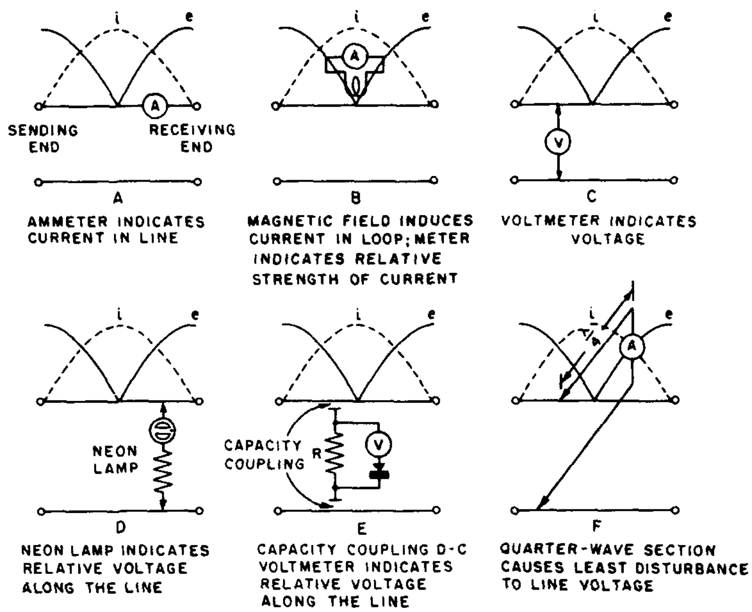


Figure 10–11.—Methods of measuring standing waves of voltage and current.

maximum current at the point where maximum current is induced in the coil—that is, at a current loop—and it indicates minimum current at a current node.

In order to determine the voltage between the lines at any point along the line, an a-c voltmeter is connected across the line, as indicated in figure 10–11, C. The voltage between the lines may also be measured by the method shown in figure 10–11, D. When the neon lamp is connected across the line the lamp will glow with a degree of brightness that is proportional to the voltage across the line at different locations along the line. If the field is strong enough, the lamp will glow when it is in close proximity to the line even if there is no physical contact. This method is convenient, although it lacks the precision of other methods. Greater precision may be obtained by the use of a sensitive type d-c

voltmeter and rectifier, as shown in figure 10–11, E. The resistor, R, is capacity-coupled to the two sides of the line, and the voltage drop across R is measured by the voltmeter.

Each of the methods described thus far has the disadvantage that a certain amount of energy is absorbed from the line. This absorption of energy at the point of measurement represents a change in impedance at this point and causes line reflections. Thus, the foregoing methods of making r-f measurements temporarily alter the normal characteristics of the line during the time the measurements are being taken.

A method of r-f measurement that causes the least disturbance to the line is shown in figure 10–11, F. It is composed of a $\frac{\lambda}{4}$ section shorted by means of an r-f ammeter. This device presents an extremely high impedance to the line, and therefore little current is needed to energize it.

When coaxial lines are used, the magnetic and electric fields are contained within the space between the outer and inner conductors and are not accessible for measurement as in open lines. In making measurements on this type of line, the arrangements shown in figure 10–12 are used. A probe is inserted in the slot, but not far enough to touch the inner conductor. The probe is a slender rod that acts as an antenna. It is excited by the electric field which is parallel to the probe. Since the line current flows parallel to the slot, the effective resistance of the coaxial line is not appreciably reduced by the presence of the slot.

Because the coupling is slight, very little energy is extracted by the probe. The r-f energy is detected by a crystal rectifier in figure 10–12, A, and the resulting d-c, which varies in magnitude with the a-c signal voltage, is amplified and fed to the voltmeter. The r-f energy is detected in figure 10–12, B, by a bolometer the resistance of which varies with temperature. This action varies the d-c current in ammeter A. The temperature of the bolometer varies with the amount of r-f current from the probe. The

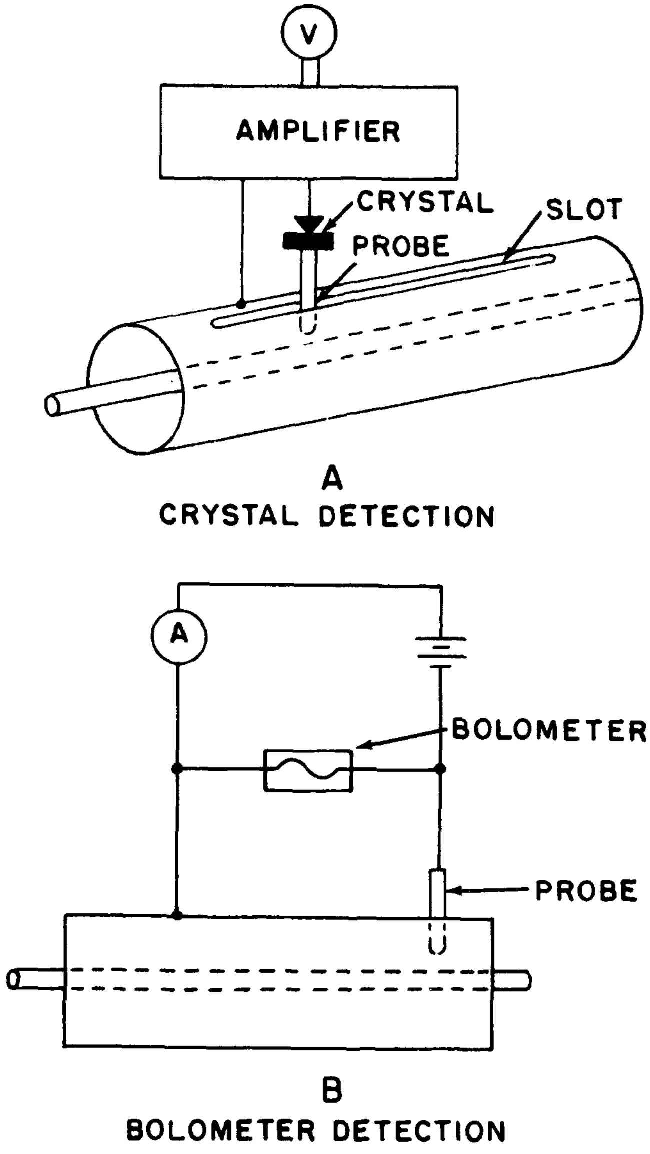


Figure 10-12.—Making measurements on coaxial lines.

bolometer itself is generally a 0.01 ampere fuse having a positive temperature coefficient.

Wavelength Measurements

Because the distance between a voltage loop and the next adjacent voltage node or a current loop and the next adjacent current node is equal to a quarter-wavelength, one wavelength is equal to four times the quarter-wave, as shown in figure 10–13.

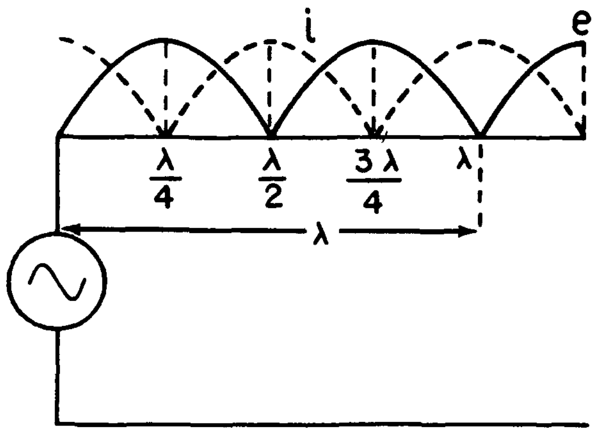


Figure 10–13.—Determination of wavelength by means of standing waves.

Because energy travels more slowly on a wire than in free space, the wavelength is a little shorter on the wire than in space. The electrical length of a wire therefore differs slightly from the length in terms of the free-space wavelength. This results from the capacitive effects between the wires and ground that decrease the velocity of propagation on the line. The spacers and insulating material used have a dielectric constant greater than air, and this also increases the effective capacitance.

The electrical quater-wavelength for various types of lines may be calculated from the formula

$$L=\frac{246 \times k}{f},$$

where L (in feet) is the quarter-wavelength, k is a constant that depends on the type of line, and f is the frequency in megacycles. The constant, k, for a parallel line is 0.975, and for an air-insulated concentric (coaxial) line is 0.85.

Lecher Lines

Lecher lines are two-wire transmission lines that are used as tuned-circuit elements or as resonant lines for the purpose of obtaining wavelength. Such lines are, in general, between ¼ and 5 wavelengths long and usually have a shorting bar that is adjustable over a considerable range of length.

In the Lecher lines shown in figure 10–14, A and B, two parallel wires are extended a distance equal to slightly more than 5 quarter-wavelengths. By the use of these lines the wavelength and the frequency of the r-f signal may be determined. The wavelength may be determined by measuring the distance between successive maxima and minima of the current or voltage waveforms. The frequency is determined by making proper substitutions in the preceding formula, transposing, and solving for f.

The use of the shorting bar and the pick-up coil to determine current maximums and minimums is illustrated by the two positions of the shorting bar shown in the figure. The current wave shown in figure 10–14, A, indicates that the standing wave of current is minimum at the location of the coupling coil. Very little current flows through the coil, and the weak magnetic field that results induces only a slight current in the pick-up coil and the indication of the ammeter is a minimum. In figure 10–14, B, the shorting bar has been moved a quarter-wavelength toward the r-f generator, and the current at the coupling coil is now maximum. Maximum current is induced in the meter pick-up coil, and the ammeter indication is a maximum.

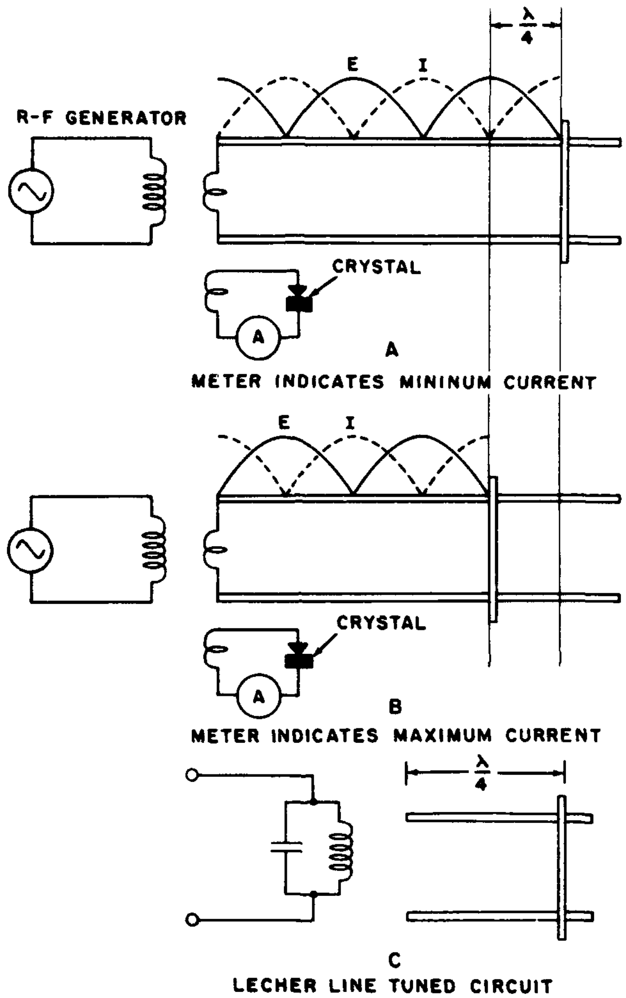


Figure 10–14.—Lecher lines.

A Lecher line one-quarter wavelength long has the characteristics of a parallel-resonant circuit (fig. 10–14, C) and therefore may be used as a tuned-plate or tuned-grid circuit

in an ultrahigh-frequency oscillator. At 400 mc a quarter-wavelength line is only a little more than 7 inches long and therefore is of a practical length for oscillators having frequencies that lie in the upper end of the v-h-f band and the lower end of the u-h-f band.

APPLICATIONS OF RESONANT LINES

In many applications, standing waves on transmission lines must be eliminated or reduced to the lowest possible level. In radar, for example, the SWR must be very near unity if satisfactory operation is to be obtained. Standing waves have the following bad effects:

1. The power-handling capacity of the line is reduced because at some points the voltage is greater than at others; and at other points the current is excessively high. At high-voltage points the insulation may break down, and at high-current points the temperature rise may be excessive.
2. The efficiency of the line is lowered because of the excessive current and accompanying I^2R loss. The line current and voltage are not in phase, hence the line power factor is low. The efficiency of transmission becomes a maximum for a given amount of power being transmitted only when the line power factor becomes unity and the effective current becomes a minimum. These conditions can exist only on a nonresonant line (no standing waves).
3. The effective resistance of the line is increased by the introduction of standing waves. There is also increased radiation loss and reduced efficiency.

For these reasons, resonant lines are seldom used to transmit large amounts of power over any considerable distance.

Resonant lines, however, have many important uses besides that of transmitting power from one point to another. For example, they may be used as metallic insulators, as wave filters and chokes, and as impedance-matching devices.

Metallic Insulators

When a quarter-wave line is shorted at the output end and is excited to resonance at the other end by the correct frequency, there are standing waves of current and voltage on the line. At the short circuit, the voltage is zero while the current is at a maximum. At the input end (the end supporting the transmission line) the current is nearly zero and the voltage is a maximum. Therefore, at the input end the $\frac{E}{I}$ ratio, and thus the impedance, is very large. Because an exceedingly high impedance across the input end looks like an insulator to the transmission line, the quarter-wave line shorted at the output end may be used as an insulator at its two open terminals (those to which the transmission line is attached).

Figure 10–15, A, shows a quarter-wave section of line acting as a stand-off insulator for a two-wire transmission line. Naturally, for direct current this section acts as a direct short on the line, but for the particular frequency that makes the section a quarter-wavelength, it acts as a highly efficient insulator. At terminals A and B there is a high voltage and a low current. Because $Z=\frac{E}{I}$, the impedance between A and B must be very high. The insulator obtains a negligible amount of energy from the line to make up any losses caused by the circulating current. If the frequency varies too widely from the value for which the section is designed, the section rapidly becomes a poor insulator and begins to act as a capacitor or inductor across the line.

Figure 10–15, B, shows a quarter-wavelength of coaxial line that is "teed" into a coaxial transmission line to support the center conductor. If the quarter-wave stubs are placed close enough together to provide adequate mechanical support, they are usually more efficient than beads of dielectric material—that is, IF THE COAXIAL LINE IS OPERATED AT ONE FREQUENCY ONLY. Metallic insulators are practical

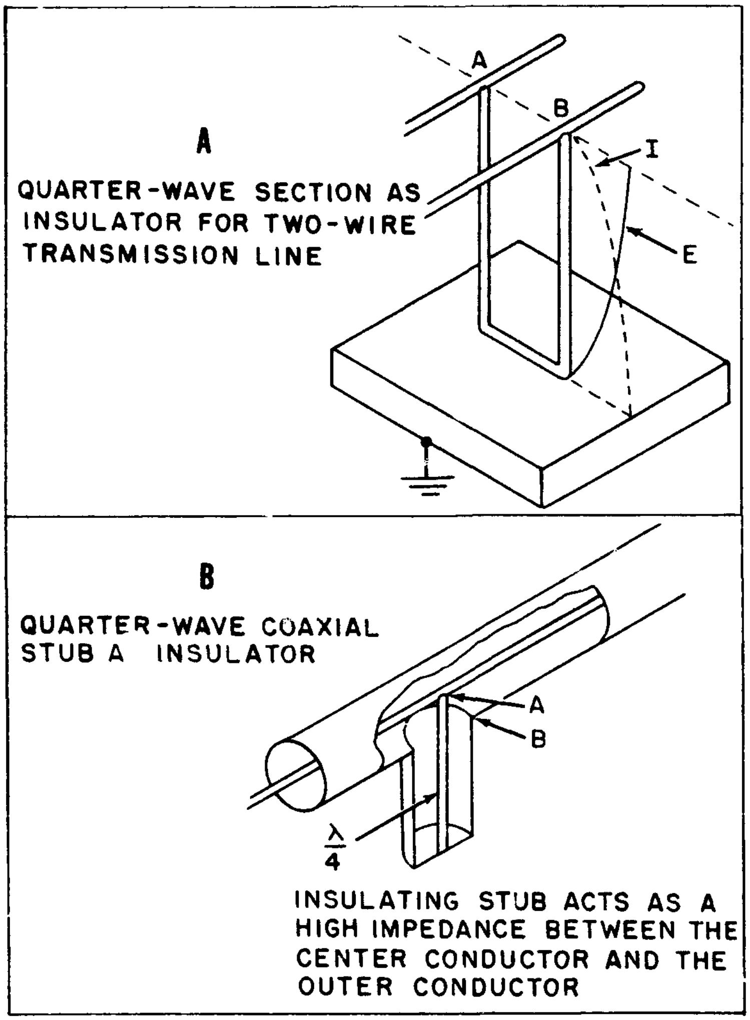


Figure 10-15.—Quarter-wave insulators.

only at the higher frequencies where the quarter-wave stub is of a practical length.

Impedance-Matching Devices

The impedance of a quarter-wave section of transmission line shorted at one end varies widely over its length, as is indicated in figure 10–16, A. At the shorted end the current is high and the voltage is low. Because $Z=\frac{E}{I}$, the impedance at the shorted end is low. At the open end, the conditions are reversed, and the impedance is high. When this section of r-f transmission line is excited, it is possible to match almost any impedance somewhere along the line. For example, a 300-ohm line may be matched to a 70-ohm line without the production of standing waves on either of the two lines that are being matched. Figure 10–16, B, shows how this connection may be made. Energy from the 300-ohm line sets up standing waves on the quarter-wave section. The connection between the 300-ohm line and the quarter-wave matching section is made at a point where the impedance of the quarter-wave line is 300 ohms. When this adjustment is made, the SWR on the 300-ohm line should be at a minimum, essentially unity. The 70-ohm line is similarly adjusted to bring about an impedance match near the shorted end.

The quarter-wave line may also be used to match a nonresonant line to a resonant line, as shown in figure 10–16, C. In order to be nonresonant, a line must be terminated in its characteristic impedance, and the terminating impedance should be approximately a pure resistance. The impedance of a shorted quarter-wave resonant section is zero at the shorting bar and increases along the line toward the open end. The shorting bar is adjusted to make a voltage maximum appear at *cd*; and the contacts, *ab*, between the nonresonant line and the quarter-wave section, are adjusted for the best match.

A half-wave section of line shorted at both ends is also used as an impedance-matching device, particularly in

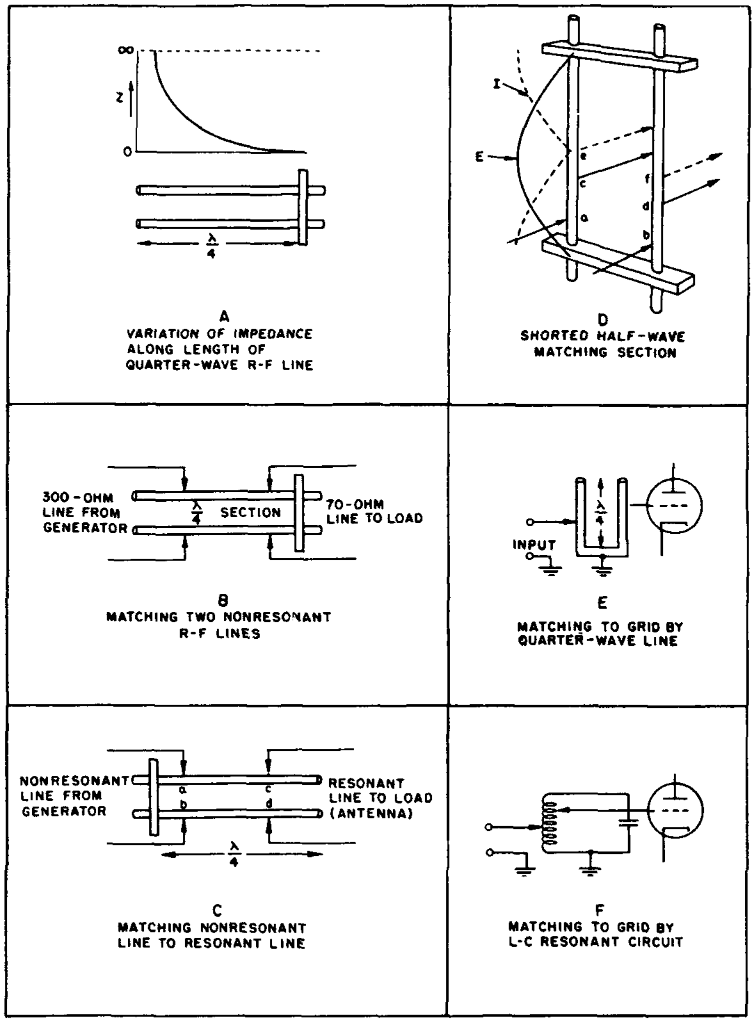


Figure 10-16.—Transmission line as an impedance-matching device.

antenna-coupling problems. Figure 10–16, D, shows a half-wave section excited at *ab* and having resonant current and voltage values as shown by the curves labeled E and I. The input (from the generator) to *ab* "sees" an impedance, Z_{ab}, equal to the $\frac{E}{I}$ ratio at that point, and the output (load) looking into *cd* "sees" a larger $\frac{E}{I}$ ratio, hence a larger impedance, Z_{cd}. The greatest impedance will be obtained at *ef* where the voltage is highest and the current lowest. Conversely, the lowest impedance points will be at the shorting bars where the current is high and the voltage low. Because the upper half of the half-wave section, or half-wave frame, repeats the impedance of the lower half, there will always be two points on the frame that have the same impedance. There will be a difference, however, in the phase of the currents involved, the current in one half being 180° out of phase with that of the other half.

Another example of the use of a shorted quarter-wave section as an impedance-matching device is shown in figure 10–16, E. In this figure a relatively low impedance input is transformed to a high impedance to match the high input impedance to a grid. The equivalent lumped circuit is shown in figure 10–16, F.

A nonshorted resonant transmission line may also be used as an impedance-matching device, as shown in figure 10–17, A. A nonshorted quarter-wave transmission line having the correct characteristic impedance may be used to match two dissimilar impedances. The necessary characteristic impedance, Z_o, of the quarter-wave matching section is

$$Z_o = \sqrt{Z_s \times Z_r},$$

where Z_s is the impedance of line 1 and Z_r is the impedance of line 2. In this figure,

$$Z_o = \sqrt{300 \times 70} = 145 \text{ (approx.)}.$$

The transformer analogy is shown in figure 10–17, B. A

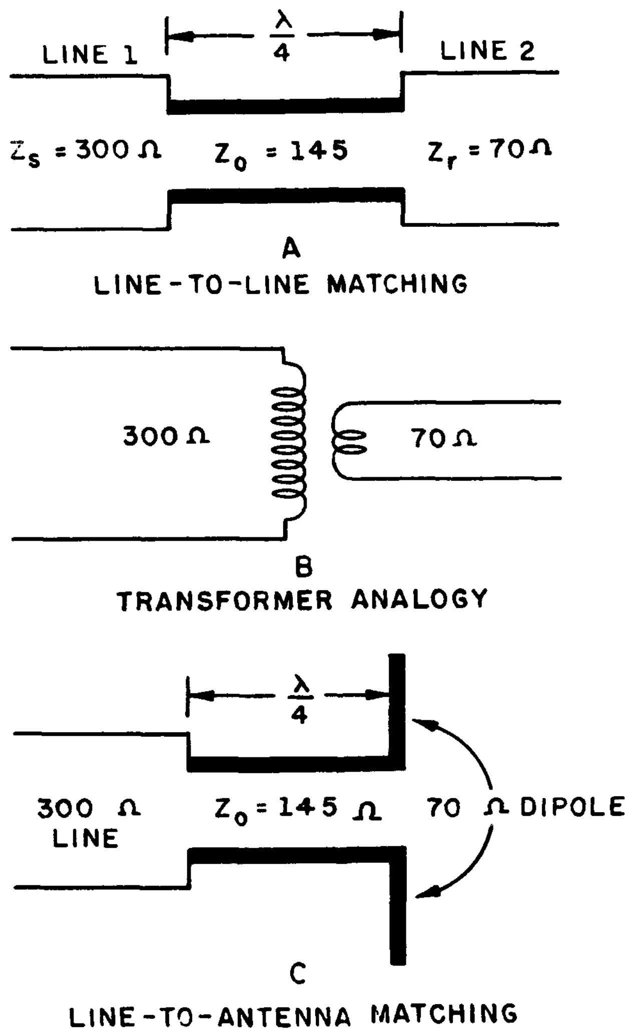


Figure 10-17.—Impedance matching with unshorted quarter-wave line.

method of connecting a 300-ohm line by means of a quarter-wave matching section is shown in figure 10–17, C.

Quarter-Wave Lines as Filters

The characteristics of a quarter-wave line allow it to be used as an efficient filter or suppressor of EVEN harmonics. Other types of filters may be used for the elimination of ODD harmonics. In fact, filters may be designed to eliminate efficiently the radiation of an entire single side band of the modulated carrier.

Suppose that a transmitter is operating on a frequency of 5 mc and it is found that the transmitter is causing excessive interference on 10 and 20 mc. In addition to the other means of eliminating radiation at these even harmonic frequencies, a resonant transmission may be used as a harmonic suppressor.

A quarter-wave line shorted at one end offers a high impedance at the unshorted end to the fundamental frequency. At a frequency twice the fundamental such a line is a half-wave line, and at a frequency four times the fundamental the line becomes a full-wave line. A half-wave or a full-wave line that is shorted at the output end offers zero impedance at its input end. Therefore, the radiation of even harmonics from the transmitting antenna can be eliminated almost completely by means of the circuit shown in figure 10–18, A.

The resonant filter line, *ab*, as shown, is a quarter-wave in length at 5 mc and offers almost infinite impedance at this frequency. In other words, the quarter-wave section looks like an insulator (to the transmission line) connected between the lines at the point where the antenna is connected. At the second harmonic, 10 mc, the line, *ab*, is a half-wave line and offers zero impedance at the antenna, thus shorting this frequency to ground. Again at 20 mc, the filter is a full-wave line and offers zero impedance. Thus, energy at this frequency is also grounded. The quarter-wave filter may be inserted anywhere along the nonresonant transmission line with similar effect—for example, at *a* in figure 10–18, B.

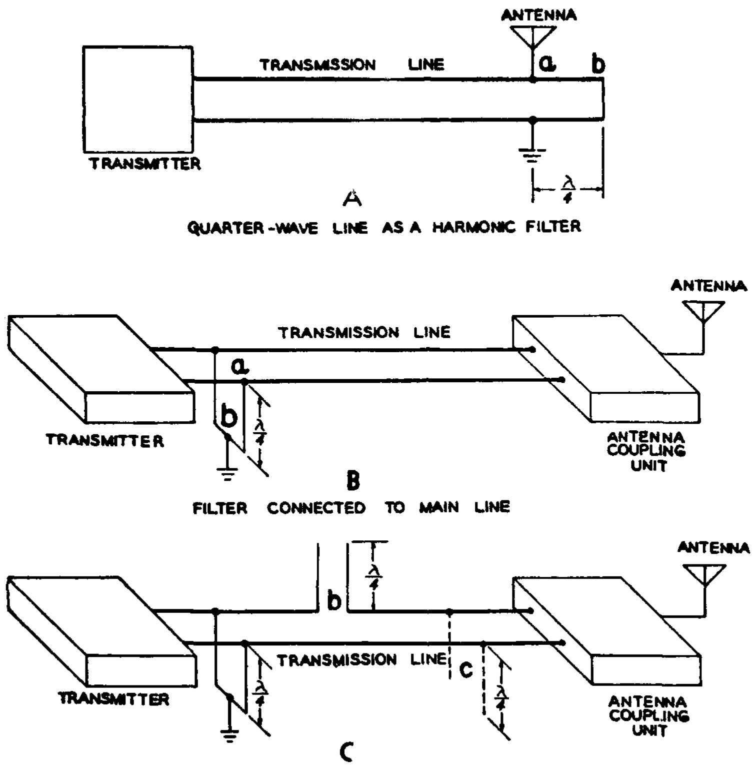


Figure 10–18.—Quarter-wave filters.

Both open and closed quarter-wave resonant lines may be used as wave filters. Figure 10–18, C, shows how more than one line filter may be connected between a transmitter and an antenna to eliminate the radiation of undesired frequencies. In this instance, a quarter-wave filter, *b*, that is open at the output end is inserted in SERIES with the transmission line. A quarter-wave line that is open at the output end offers low impedance at the input end to the fundamental frequency. At each odd harmonic such a line is an odd multiple of a quarter-wave and therefore offers little im-

pedance to odd harmonics. Actually, at the fundamental and odd harmonics the impedance at *b* is so low that it may be considered a continuous line, as if a short were placed across the base of the quarter-wave line. Thus, the quarter-wave open-filter line, *b*, in figure 10–18, C, passes the fundamental and odd harmonics along the line to the antenna-coupling unit. At even harmonics, however, the length of the open line (at *b*) becomes a half wave, or some multiple of a half wave, so that line *b* offers high impedance to the even harmonics and blocks their passage to the antenna-coupling unit.

Unfortunately, the foregoing methods of inserting wave filters in shunt with a line cannot be used to eliminate odd harmonics, because any attempt to eliminate the odd harmonics also results in serious loss to the fundamental frequency. For example, assume that line *c* of figure 10–18, C, is a quarter-wave at the third harmonic (15 mc). This frequency would be eliminated effectively before it could reach the antenna-coupling unit. However, the fundamental that is to be transmitted would also be greatly attenuated. If a line is a quarter-wave in length at 15 mc it is a twelfth-wave in length at 5 mc (wavelength varies inversely with frequency). A line a twelfth-wave in length would act as a capacitor and offer a low impedance to 5 mc. Therefore, although 15-mc radiation would be suppressed, the desired carrier would also be suppressed considerably.

QUIZ

1. What are four uses of resonant r-f lines other than for the transmission of power?
2. What is the effect on the characteristic impedance of a two-wire line if the wires are moved farther apart?
3. What is the phase relation between voltage and current on a line of infinite length?
4. Why do the waveforms diminish in amplitude along a line of infinite length?
5. What is the constant ratio of voltage to current called on a line that is terminated in an impedance equal to this ratio?
6. What is the relative magnitude of the rms voltage at the open end of a transmission line that is one wavelength long (fig. 10–3, B)?
7. What is the relative magnitude of the rms current at the open end of a transmission line that is one wavelength long (fig. 10–3, C)?
8. What is the relative magnitude of the rms current at the shorted end of a transmission line that is one wavelength long (fig. 10–4)?
9. What is the relative magnitude of the rms voltage at the shorted end of a transmission line that is one wavelength long (fig. 10–4)?
10. What is the relative magnitude of the load impedance compared with the characteristic impedance of a nonresonant line?
11. In an open-end resonant line, why is the impedance prevented from being zero at odd quarter-wavelengths from the terminal end of the line?
12. What type of circuit would a generator "see" if it is connected one-half wavelength from the end of an open-end resonant line?
13. What type of circuit would a generator "see" if it is connected three-eighths wavelength from the end of an open-end resonant line?
14. A shorted transmission line one-half wavelength long acts like what kind of circuit?
15. When a transmission line is terminated in a capacitive reactance equal to the characteristic impedance of the line, how are the voltage and current distributions affected?
16. Express the SWR in terms of (1) voltage, (2) current, and (3) impedance.
17. What effect does increasing the mismatch between line and load have on the SWR?
18. What is the principal disadvantage of parallel-wire transmission lines?

19. The twisted-pair transmission line is not used for high frequencies because of what type of losses?
20. What is the principal advantage of the shielded-pair transmission line?
21. May higher or lower frequencies than the cutoff frequency be transmitted by a hollow waveguide?
22. What places a limit on the narrow width of a rectangular waveguide?
23. Why are waveguides not practical at the lower frequencies?
24. Define the TE mode of operating a waveguide.
25. Define the TM mode of operating a waveguide.
26. What is meant by the dominant mode of a waveguide?
27. As the $\frac{\lambda}{4}$ section of figure 10–11, F, is moved along the transmission line, current loops are indicated by what relative value of current on the ammeter (maximum or minimum)?
28. How are voltage measurements made on coaxial lines?
29. What are two of the uses of Lecher lines?
30. Give three of the bad effects of standing waves.
31. What is the relative magnitude of the impedance at the input end of a shorted quarter-wave line used as a metallic insulator?
32. Why are metallic insulators not practical at the lower frequencies?
33. Why may a wide variety of impedances be matched by means of a quarter-wave line shorted at one end?

CHAPTER 11

ANTENNAS AND PROPAGATION

PRINCIPLES OF RADIATION

Fundamental Concepts

A radio-frequency current flowing in a wire of finite length can produce electromagnetic fields that may be disengaged from the wire and set free in space. The principles of the radiation of electromagnetic energy are based on the laws that a MOVING ELECTRIC FIELD CREATES A MAGNETIC FIELD and conversely, a moving magnetic field creates an electric field. The created field (either electric or magnetic) at any instant is in phase in time with its parent field, but is perpendicular to it in space. THESE LAWS HOLD TRUE WHETHER OR NOT A CONDUCTOR IS PRESENT.

The electric (E) and magnetic (H) fields are perpendicular to each other and perpendicular to the direction of motion through space. A right-hand rule may be applied that relates the directions of the E field, the H field, and the propagation. This rule states that if the thumb, forefinger, and middle finger of the right hand are extended so that they are mutually perpendicular, the thumb will point in the direction of the electric field, the forefinger in the direction of the magnetic field, and the middle finger in the direction of propagation.

In the instantaneous cross section of a radio wave shown in figure 11–1 the E lines represent the electric field and the

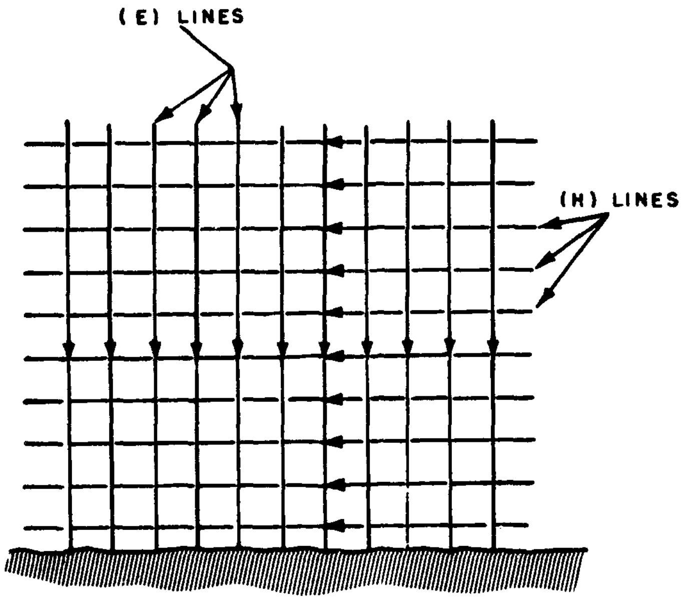


Figure 11-1.—Instantaneous cross section of a radio wave.

H lines represent the magnetic field. If the right-hand rule is applied, the thumb points downward, representing the direction of the E lines; the forefinger, to the left, representing the direction of the H lines; and the middle finger away from the observer, representing the direction of propagation.

When r-f current flows through a transmitting antenna, radio waves are radiated from it in all directions in much the same way that waves travel on the surface of a pond into which a rock has been thrown. It has been found that these radio waves travel at a speed of approximately 186,000 miles per second (300 million meters per second). The frequency of the radio wave radiated by the antenna will be equal to the frequency of the r-f current.

Since the velocity of the radio wave is constant regardless of its frequency, to find the wavelength (which is the distance traveled by the radio wave in the time required for one cycle)

it is necessary only to divide the velocity by the frequency of the wave—

$$\lambda=\frac{300,000,000}{f},$$

where λ is the distance in meters from the crest of one wave to the crest of the next, f the frequency in cycles per second, and 300,000,000 the velocity of the radio wave in meters per second. This relationship is important in radio communications. It can also be expressed as

$$f=\frac{300}{\lambda},$$

where f is in megacycles, λ is in meters, and 300 is the velocity of propagation of the radio wave in millions of meters per second.

For example, the frequency of the current in a transmitting antenna that is radiating an electromagnetic wave having a wavelength of 2 meters is $\frac{300}{2}$, or 150 megacycles. Radio waves are usually referred to in terms of their frequency and are discussed in considerable detail later in this chapter.

In figure 11–2 a piece of wire is cut in half and each half is attached to the terminals of a high-frequency a-c generator. The frequency of the generator output is chosen so that each half of the wire is one-quarter of the wavelength, $\frac{\lambda}{4}$, corresponding to the generator frequency. The result is a common type of antenna known as a DIPOLE and is shown in figure 11–2, A.

At a given instant, the right-hand terminal of the generator is positive and the left-hand terminal is negative. Since like charges repel, electrons will flow away from the negative terminal as far as possible, while the positive terminal will attract electrons to it. Figure 11–2, B, shows the direction and distribution of electron flow at this instant. The current distribution curve indicates that the current flow is

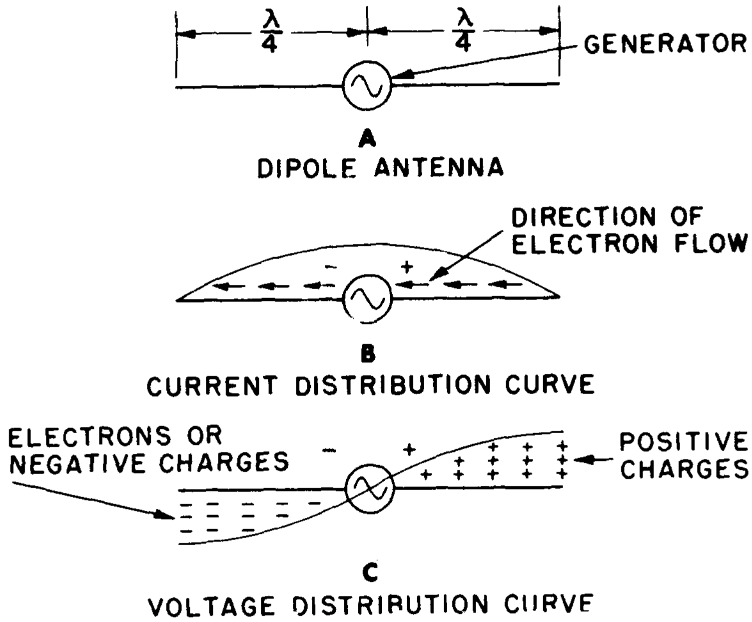


Figure 11-2.—Dipole antenna showing current and voltage distribution.

greatest at the center of the dipole and zero at the ends. At any given point along the antenna, except at the ends, the current variation is assumed to be sinusoidal with respect to time (the generator voltage has sine waveform). The relative current distribution is also sinusoidal with respect to the antenna length. Thus an r-f ammeter inserted near the center of the antenna will indicate a relatively large effective current and one inserted near the end will indicate a small effective current. The relative current distribution over the antenna will always be the same no matter how much or how little current is flowing, but the current amplitude at any given point on the antenna will vary directly with the amount of voltage developed at the generator terminals.

The generator voltage initiates the flow of antenna current. The action of the antenna is partly like that of a capacitor. When a capacitor becomes fully charged its voltage is maxi-

mum and the charging current ceases. In figure 11–2, C, the antenna voltage near the ends is maximum at the instant that the charging current is zero. Although no current flows at this instant there is a maximum accumulation of electrons at the left end of the antenna and a deficit at the right end. Most of the charges are at the ends trying to get as far from the generator terminals as possible (like charges repel). The antenna voltage, like the antenna current, varies sinusoidally with respect to time. Also the antenna voltage varies sinusoidally with respect to the antenna length. Thus an r-f voltmeter connected between ground and one end of the antenna indicates a relatively large effective (rms) voltage. As the end probe is moved toward the antenna center the effective voltage is decreased to a low value. The antenna has both distributed inductance and capacitance and acts like a resonant circuit. At the center the current and voltage are in phase with each other; in the antenna wire between the center and the ends they are out of phase.

Summarizing:

1. A current having sine waveform flows in the antenna. Its distribution is sinusoidal, as shown in figure 11–2, B.
2. A sinusoidal distribution of charge, as shown in figure 11–2, C, exists on the antenna. Every half cycle the charges reverse position.
3. The sinusoidal variation in charge (voltage) is out of phase with the sinusoidal variation in current by one-quarter of a cycle, or 90°, except at the center and the ends where the current and voltage are in phase.

Induction Field

An alternating current flows in the antenna; therefore an alternating magnetic field, H, is set up around the antenna as shown at one instant in figure 11–3, A. Alternate positive and negative charges also appear on the antenna, causing an electric field (E in fig. 11–3, B) to be set up. This field is represented by lines of force drawn between the positive and negative charges (fig. 11–3, B). The arrow heads indicate the direction a unit positive charge would move at those

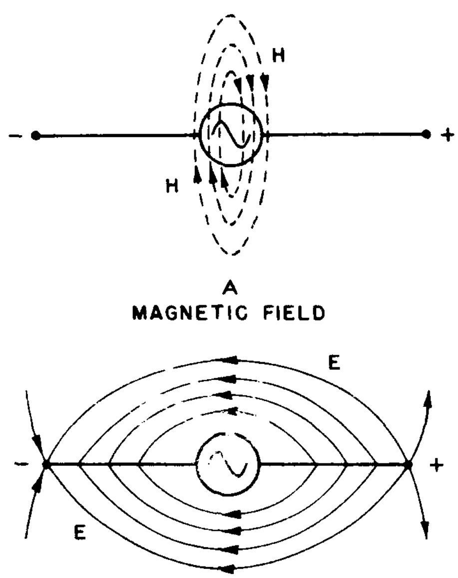


Figure 11-3.—Instantaneous field around an antenna.

points. Because the current and voltage that produce these fields are 90° out of phase the two fields must also be out of phase by 90°. Thus in spite of the fact that they are mutually perpendicular, these fields do not constitute the radiated electromagnetic field that passes through space from the transmitting antenna to the receiving antenna.

On the other hand, the magnetic and electric components of the radiated field are in phase with each other. The energy contained in the induction field cannot be detached from the antenna. The amplitude of the induction field energy varies inversely as the square of the distance from the antenna, and consequently its effect is entirely local. However, its effect must be considered in making field strength measurements of the radiation field in the vicinity of the antenna—that is, if only the field strength of the

radiation field is to be measured. At $\frac{\lambda}{2\pi}$ wavelengths away from the antenna the field strengths of the two fields are equal. This distance is approximately one-sixth wavelength. At distances of a few wavelengths away from the antenna the induction field becomes negligible.

Radiation Field

Although both a magnetic field and an electric field are radiated into space simultaneously, only the electric field is considered at present. The charges producing the electric field are constantly moving from one end of the antenna to the other as the polarity of the voltage at the generator changes. At one instant, one end of the antenna is positive; an instant later the antenna is uncharged. A negative charge next appears where the positive charge was. Then the antenna is again uncharged, and the cycle repeats.

In figure 11–4, A, electric flux lines are drawn between positive and negative charges. An instant later (fig. 11–4,

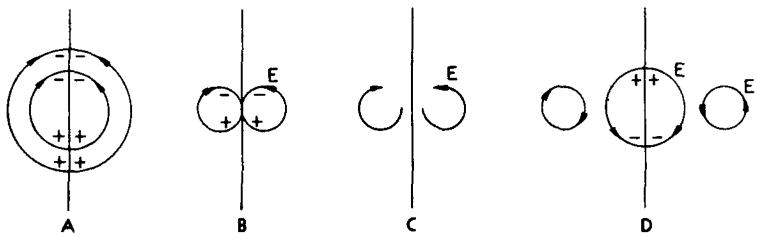


Figure 11–4.—Creation of closed electric flux lines on a half-wave antenna.

B), the antenna is nearly discharged as the charges approach each other, thus bringing together the two ends of the flux lines associated with them. When the charges do touch, they seem to disappear, and their flux lines should also disappear. Most of the flux lines that represent the INDUCTION FIELD do disappear, but some flux is repelled by other lines nearer the antenna and, as in figure 11–4, C, the

repelled flux lines are left with their heads touching their tails. A closed electric field is thus created without an associated electric charge.

An instant after the independent field has been formed, the antenna is charged again in the opposite direction and produces lines of force that repel the recently formed independent electric field. Figure 11-4, D, shows that the repelling field is of the proper polarity to do this. The radiated field is forced away from the antenna at the speed of light.

As previously stated, a moving electric field generates a perpendicular magnetic field in phase with it. Therefore, because the radiated electric field is moving, it generates a magnetic field in accordance with this principle. The result is a radiated electromagnetic field that can travel great distances and deliver a usable part of its energy to a receiving antenna.

In the preceding discussions, the magnetic field generated by the antenna current has been ignored as a factor in generating the radiated field, but, by similar reasoning, magnetic lines of force may become detached from the antenna. Because the detached lines move away from the antenna, they generate a perpendicular in-phase electric field. The result is also a radiated electromagnetic field.

The electromagnetic radiation from the antenna is thus apparently made up of two components—the electric generated field and the magnetic generated field. These two fields can be shown to add and give a single sinusoidally varying radiated field.

The strength of the radiated field varies inversely with the distance.

Reception

If a radiated electromagnetic field passes through a conductor, some of the energy in the field will set electrons in motion in the conductor. This electron flow constitutes a current that varies in accordance with the variations of the field. Thus, a variation of the current in a radiating

antenna causes a similar varying current (of much smaller amplitude) in a conductor at a distant location. Any intelligence being produced as current in a transmitting antenna will be reproduced as current in a receiving antenna. The characteristics of receiving and transmitting antennas are similar, so that a good transmitting antenna is also a good receiving antenna.

BASIC ANTENNA PRINCIPLES

General

An antenna is a conductor or system of conductors that serves to radiate or intercept energy in the form of electromagnetic waves. In its elementary form an antenna, or aerial, may be simply a length of elevated wire like the common receiving antenna for an ordinary broadcast receiver. However, for communication and radar work, other factors make the design of an antenna system a more complex problem. For instance, the height of the radiator above ground, the conductivity of the earth below it, and the shape and dimensions of an antenna all affect the radiated-field pattern in space. Also, the antenna radiation often must be directed between certain angles in either the horizontal or the vertical plane, or both.

An antenna may be constructed to resemble a resonant two-wire line with the wires so arranged that the fields produced by the currents in the wires add in some directions instead of canceling completely. Figure 11–5, A, shows one way to prevent cancellation of the fields by making the earth one conductor. This permits considerable separation of the conductors. In this manner the fields resulting from the current expand considerably farther into space than if the other conductor were nearby, and therefore can be detached from the radiating conductor by rapid reversals much more easily. Another way to accomplish the radiation is to spread the ends of the two-wire line as shown in figure 11–5, B, 180° (as shown in fig. 11–5, C). The currents, which canceled each other's fields in figures 11–5, B, now aid in pro-

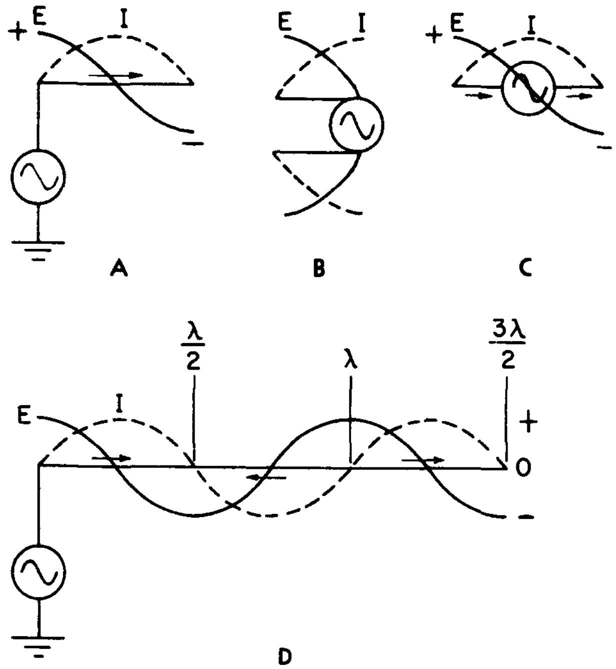


Figure 11-5.—Half-wave and multiple half-wave antennas.

ducing a field in space (fig. 11–5, C) similar to that produced in figure 11–5, A.

The antenna shown in figure 11–5, A, can be extended, as shown in figure 11–5, D. Current flowing to the right is represented by the positive portion of the current curve, and current flowing to the left (reflected from the right-hand end of the wire) is represented by the negative portion of the curve. Similarly, voltages at any point on the antenna are positive or negative with respect to ground according to the position of the voltage curve above or below the axis represented by the antenna. The effectiveness of such an antenna is not greatly increased by extending it horizontally close to the earth because currents flowing in opposite directions side by side produce canceling fields in

some directions. However, if the antenna extends vertically above the earth, it is possible to elevate the effective radiation field a greater distance by operating the antenna at some harmonic, such as the third, fifth, or seventh harmonic of the fundamental frequency. The result is that the intensity of the radiated field at various points in space is considerably changed when compared with the field of the simple dipole.

Nonresonant lines also can be expanded to antennas, but they are not efficient radiators. Resonant conductors are more efficient radiators because they have large standing waves of voltage and current, and hence they produce intense fields with a minimum of generator current and voltage. Thus the antenna shown in figure 11–5, A, which is cut to an electrical half wavelength, also radiates other frequencies, but its effectiveness as a radiator diminishes as the standing waves of current and voltage decrease.

Electrical Length

If an antenna is made of very small wire and is isolated perfectly in space, its electrical length corresponds closely to its physical length. Thus, in free space, a 1-wavelength antenna for 10 meters would be 10 meters in length, and a half-wave length antenna for the same signal would be 5 meters in length. In actual practice, however, the antenna is never isolated completely from surrounding objects. For example, the antenna will be supported by insulators whose dielectric constant is greater than 1. Therefore the velocity of the wave along the conductor is always slightly less than the velocity in space, and the physical length of the antenna will be correspondingly less (by about 5 percent) than the corresponding wavelength in space. The physical length, L, in feet, of a half-wave antenna for a given frequency is derived as follows:

Since

$$\lambda=\frac{300}{f} \text{ and } \frac{\lambda}{2}=\frac{300}{2f},$$

$$L=\frac{300\times 3.26\times 0.95}{2f}=\frac{468}{f},$$

where f is the frequency in megacycles, 3.26 feet equal 1 meter, and 0.95 represents the velocity of the wave in the antenna compared to that in free space. This formula does not apply to antennas longer than one-half wavelength.

Antenna Input Impedance

The antenna input impedance determines the antenna current at the feed point for a given value of r-f voltage at that point. The input impedance may be expressed mathematically by Ohm's law for alternating current—

$$Z=\frac{E}{I},$$

where Z is the antenna impedance and E and I are the r-f voltage and current respectively. Impedance is also expressed as

$$Z=R\pm jX,$$

where R and X are the input resistance and reactance respectively.

In a half-wave antenna, the current is a maximum at the center and zero at the ends; whereas the voltage is a maximum at the ends and minimum at the center. The impedance, therefore, varies along the antenna and is minimum at the center and a maximum at the ends. Thus, if energy is fed to a half-wave antenna at its center, it is said to be CENTER FED (current fed); if energy is fed at the ends it is said to be END FED (voltage fed). In the case of a half-wave antenna isolated in free space, the impedance is approximately 73 ohms at the center and 2,500 ohms (allowing for losses) at the ends. The intermediate points have intermediate values of impedance.

Figure 11–6, A, is a plot of the input resistance of center-fed antennas for various wavelengths. Resistance values

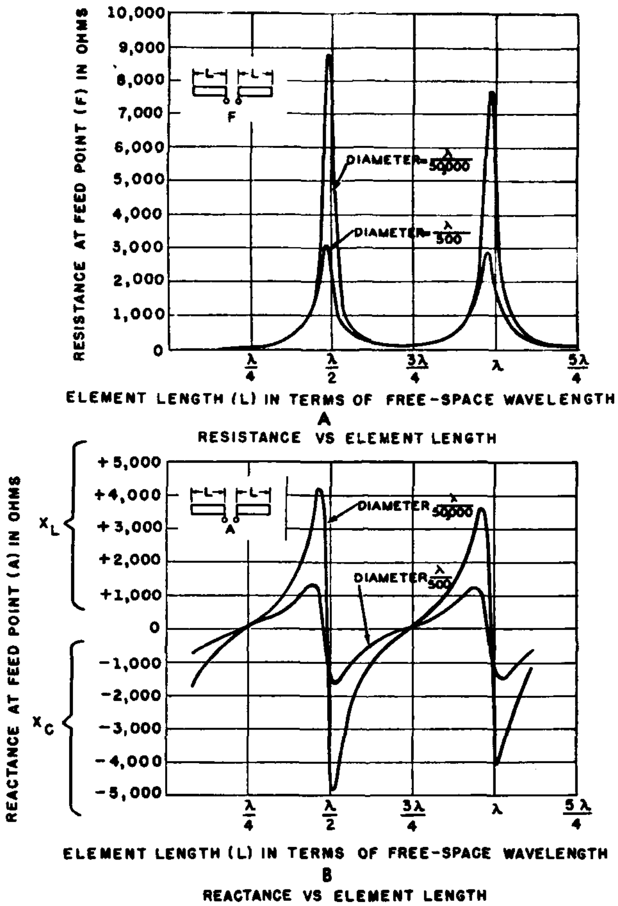


Figure 11-6.—Impedance curves for a center-fed antenna.

for both a thin and a thick antenna are plotted so that the effect of the diameter of the wire is apparent. In figure 11–6, B, the reactance is plotted as a function of wavelength. The curves show that an antenna may be either inductive or capacitive depending on its length, and that abrupt changes of impedance occur at multiples of a half-wavelength. The points in figure 11–6, B, where the reactance

curves cross zero indicate the resonant lengths of the antenna. Because the curves are plotted in terms of the free-space wavelength, the effect of the reduced velocity of the wave motion along the antenna is shown by the curves. For example, a half-wave antenna element is resonant only when it is less than the free-space half-wavelength. This foreshortening is caused by the increased capacitance associated with the elements. If the diameter of the radiator is large, for example $\frac{\lambda}{500}$, the increased capacitance is greater than for a thin element. As a result, the large-diameter radiator is foreshortened more than the thin radiator.

Figure 11-6 may be used to calculate the input impedance of center-fed antennas. For example, let it be required to find the impedance of a thin $\left(\text{diameter}=\frac{2\lambda}{50{,}000}\right)$ antenna whose half length is five-eighths of the wavelength being fed to it. In this case the antenna is not fully resonant. The impedance includes both resistance and reactance. The resistance is located on the proper curve halfway between $\frac{\lambda}{2}$ and $\frac{3\lambda}{4}$ and is approximately 150 ohms. Similarly, the reactance is found to be capacitive and approximately 1,100 ohms. The impedance in ohms is,

$$Z=150-j1{,}100=1{,}110\ \angle-82.2°.$$

Thus, for maximum transfer of energy to the antenna a feedline to a $\frac{5\lambda}{8}$ center-fed antenna in free space must be designed to present a conjugate impedance of $150+j1{,}100=1{,}110\ \angle+82.2°$. In this case the feedline has a resistance of 150 ohms and an inductive reactance of 1,100 ohms.

The input impedance of an antenna is affected by the presence of nearby conductors (for example, the rigging on ships). Any object that can be affected by the induction field will distort the field and also the antenna voltage and current distribution. Therefore, the input impedance will be changed, and necessary corrections must be made to obtain the best match to each antenna. Because this effect

is almost always difficult if not impossible to calculate, corrections are usually determined empirically by trial-and-error methods.

Radiation Resistance

The antenna at the end of the transmission line is equivalent to a resistance that absorbs a certain amount of energy from the generator. Neglecting the losses that occur in the antenna, this is the energy that is radiated into space. The value of resistance that would dissipate the same power that the antenna dissipates is called the RADIATION RESISTANCE of the particular antenna. The power dissipated in a resistor is equal to I^2R. Likewise, the power dissipated in (radiated from) and antenna is equal to the current (at the feed point) squared times the radiation resistance of the antenna.

Figure 11–7 shows how the radiation resistance varies

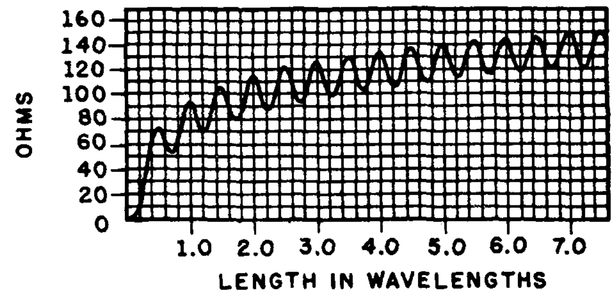


Figure 11–7.—Radiation resistance of antennas in free space plotted against length.

with antenna length, for an antenna in free space. For a half-wave antenna the radiation resistance is approximately 73.2 ohms, measured at the current maximum, which is at the center of the antenna. For a quarter-wave antenna the radiation resistance measured at the current maximum is approximately 36.6 ohms. The radiation resistance is also affected somewhat by the height of the antenna above ground and by its proximity to nearby objects. Other small antenna losses are caused by the ohmic resistance of the conductor, corona discharge, and insulator losses.

Wave Polarization

The position of a simple antenna in space determines the polarization of the emitted wave; that is, the direction of the electric lines of force determines the polarization of the wave. Thus, an antenna that is vertical with respect to the earth radiates a vertically polarized wave, while a horizontal antenna radiates a horizontally polarized wave. Figure 11–8, A, shows the vertical electric field component of a vertical antenna as a sine wave in the plane of the paper. Figure 11–8, B, shows the horizontal electric field component of a

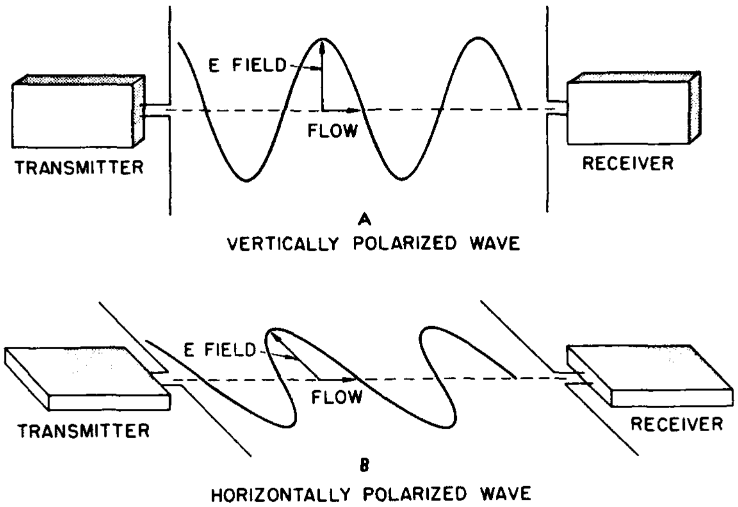


Figure 11–8.—Vertical and horizontal polarization.

horizontal antenna as a sine wave lying in a horizontal plane. The first wave is vertically polarized; the second, horizontally polarized. For low frequencies the polarization is not disturbed and the radiation field has the same polarization at the distant receiving station that it had at the transmitting antenna. At high frequencies, however, the polarization usually varies, sometimes quite rapidly, because the wave splits into several components which follow different paths. These paths will not be the same length; therefore the recom-

bined electric vectors representing the several components generally will not be parallel. If this is the case, the path traced by the point of the resultant vector may be circular or elliptical, and such a radiated field is known as either a CIRCULARLY or an ELLIPTICALLY POLARIZED FIELD.

When the antennas are close to the ground, vertically polarized waves yield a stronger signal close to the earth than do horizontally polarized waves. However, when the transmitting and receiving antennas are at least 1 wavelength above ground, the two types of polarization give approximately the same field intensities near the surface of the earth. When the transmitting antenna is several wavelengths above ground, horizontally polarized waves result in a stronger signal close to the earth than is possible with vertical polarization.

Polar Diagrams

The variation of signal strength around an antenna can be shown graphically by polar diagrams as in figure 11–9. Zero distance is assumed to be at the center of the chart (indicating the center of the antenna) and the circumference of the tangent circles is laid off in angular degrees. Computed or measured values of field strength then may be plotted radially in a manner that shows both magnitude and direction for a given distance from the antenna. Field strengths in the vertical plane are plotted on a semicircular polar chart (not shown in the figure) and are referred to as vertical polar diagrams.

BASIC TYPES OF ANTENNAS

Hertz Antenna

Any antenna that is one-half wavelength long, or any even or odd multiple thereof, is a Hertz antenna and may be mounted either vertically or horizontally. A distinguishing feature of all Hertz antennas is that they need not be connected conductively to the ground, as are other antennas to be described. At the low and medium frequencies these

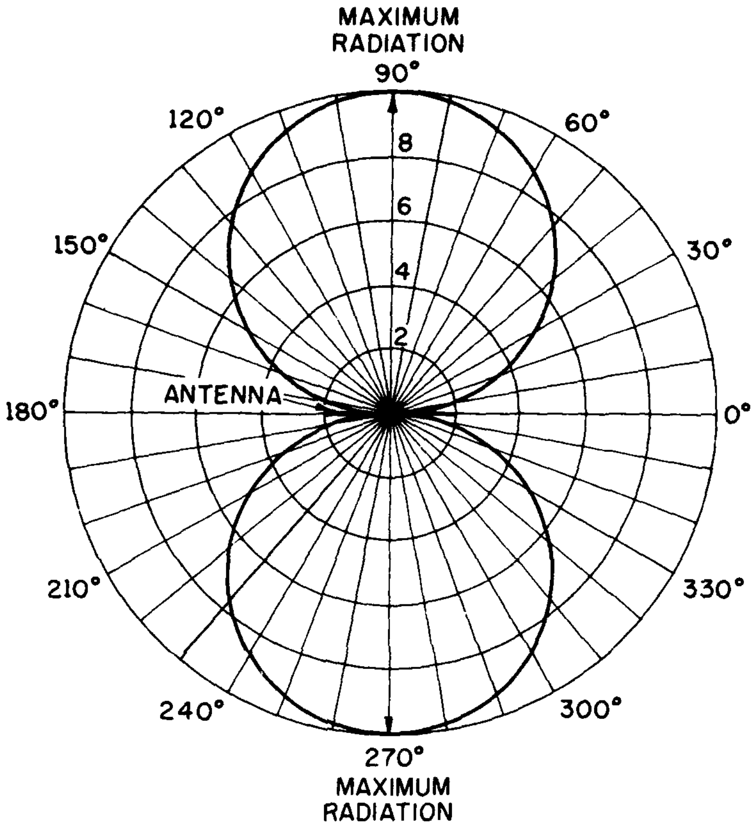


Figure 11-9.—Polar diagram of an antenna showing relative field strength.

antennas are rather long and have little use in the Navy except at shore stations where there is room for them. Vertical half-wave and five-eighths wave antennas are widely used with a-m broadcasting stations and have been built to heights of 1,000 feet or more for the lower broadcast frequencies. At the medium and high frequencies they are used extensively in fixed service when operation is not required at a large number of frequencies. This type of antenna is not

particularly suited to services where a large number of different and unrelated frequencies must be transmitted using the same antenna, such as aboard ship.

Half-wave antennas showing two different methods of connecting the feedline together with the equivalent resonant circuits are shown in figure 11-10. For a half-wave dipole,

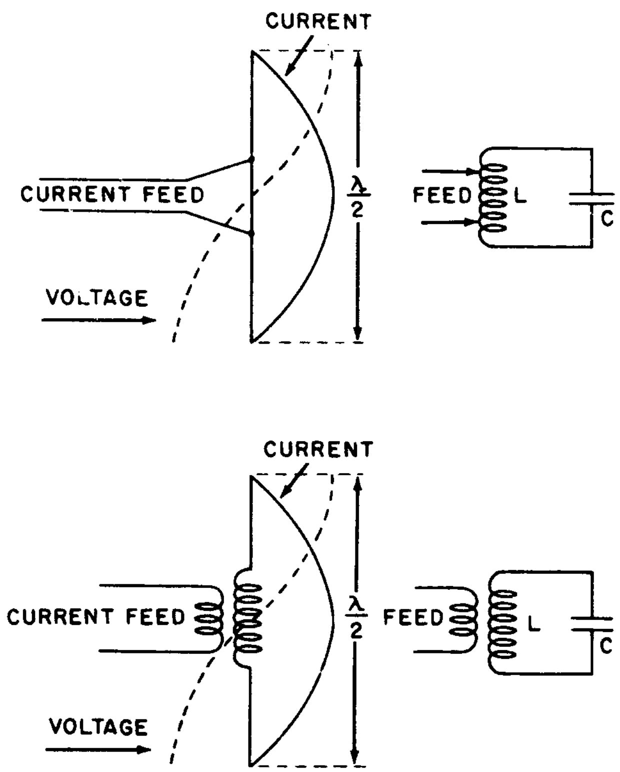


Figure 11-10.—Hertz antennas and equivalent circuits.

the effective current is maximum at the center and minimum at the ends, while the effective voltage is minimum at the center and maximum at the ends. The voltage and current relationships are similar to those of the simple dipoles shown in figure 11-10.

Marconi Antenna

A grounded antenna whose length is one-fourth wave or any odd multiple thereof is known as a Marconi antenna. Figure 11–11 illustrates the principle of a Marconi antenna

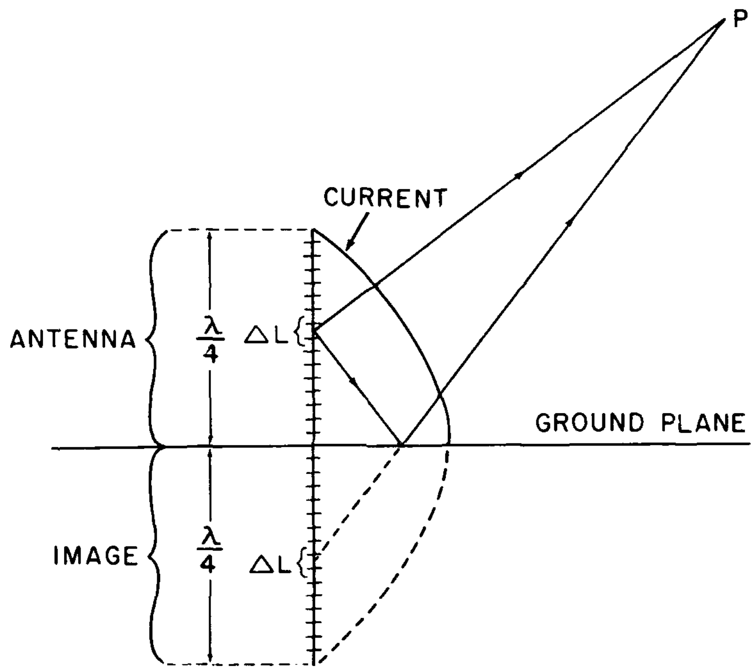


Figure 11–11.—Marconi antenna and image.

mounted on the surface of the earth. The transmitter may be connected between the bottom of the antenna and the earth. Although the antenna is only one-quarter wavelength long, the reflection in the earth is equivalent to another quarter-wave antenna. By this arrangement half-wave operation can be obtained from an antenna only one-quarter wavelength long. The impedance, voltage, and current relationships are similar to those in a half-wave antenna except that the input impedance at the base of a Marconi

antenna is 36.6 ohms; whereas the input impedance of a Hertz antenna at the center is 73.2 ohms.

The quarter-wave antenna is used extensively with portable transmitters. On an airplane a quarter-wave mast or trailing wire is the antenna, and the fuselage produces the image. Similar installations are made on ships. A quarter-wave mast or horizontal wire is the antenna, and the hull and superstructure provide the image.

The effective current in the Marconi quarter-wave grounded antenna is maximum at the base and minimum at the top. The voltage is maximum at the top and minimum at the base.

ANTENNA TUNING

Aboard ship, antennas used for communications at the medium frequencies are not usually of the proper length to give optimum performance at the operating frequency. This condition exists because these antennas are all of standard size and shape or are installed in whatever space may be available for them and because they are each operated at more than one frequency. All equipment must be able to operate at any frequency within its tuning range. In this case, then, it is necessary to employ some means at the transmitter to adjust the antenna for reasonable efficiency at any frequency regardless of the physical dimensions or arrangement of any antennas that might be available.

Since each transmitter is usually associated with only one antenna, which is of fixed length, the adjustment of the effective length of the antenna must be made by electrical means. This process is called ANTENNA TUNING and is accomplished by adding either inductance or capacitance to the antenna at the point where it is fed from the transmitter or transmission line, as shown in figure 11–12. Added inductance has the property of increasing the effective electrical length, while capacitance decreases it. In this manner the antenna can be made to respond as if it contained a whole number of quarter waves along its length.

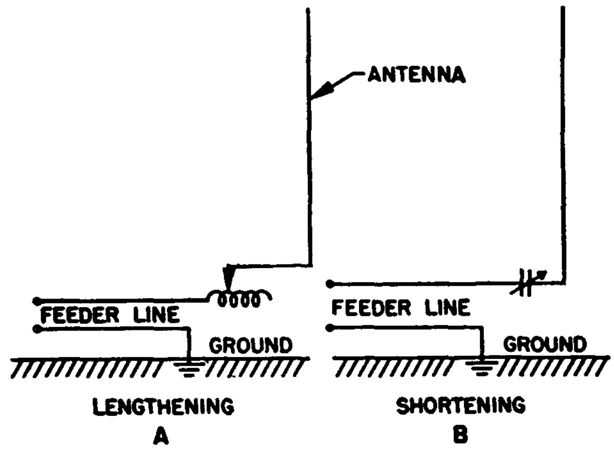


Figure 11-12.—Methods of correcting the electrical length of a grounded antenna.

By tuning the antenna properly, the standing waves are increased and the radiated energy is increased.

Sometimes, particularly at low frequencies, it is not desirable to make the quarter-wave grounded antenna the full physical quarter-wave in height shown in figure 11-13, A. Instead, it may be made shorter physically and then made the correct length electrically by top-loading it with a series inductor (fig. 11-13, B) or a parallel capacitor (fig. 11-13, C).

If the antenna is slightly more than a quarter wavelength high, the input at the base will be inductive, requiring the addition of a capacitor in series with the feed to bring the antenna into resonance.

RADIATION PATTERN FOR HALF-WAVE ANTENNAS

Since the current is greatest at the center of a dipole, maximum radiation takes place at this point and practically no radiation takes place from the ends. If this antenna could be isolated completely in free space, the points of maximum radiation would be in a plane perpendicular to the

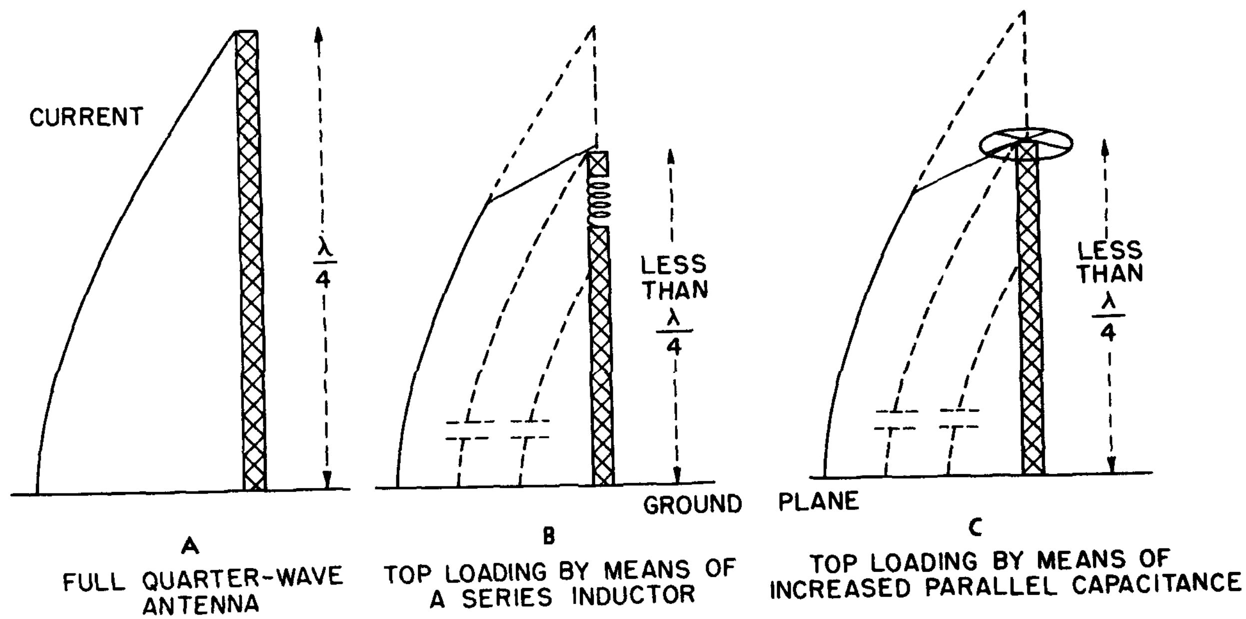


Figure 11-13.—Vertical quarter-wave antennas and methods of loading.

plane of the antenna at its center. The doughnut-shaped surface pattern is shown in figure 11-14, A, and the horizon-

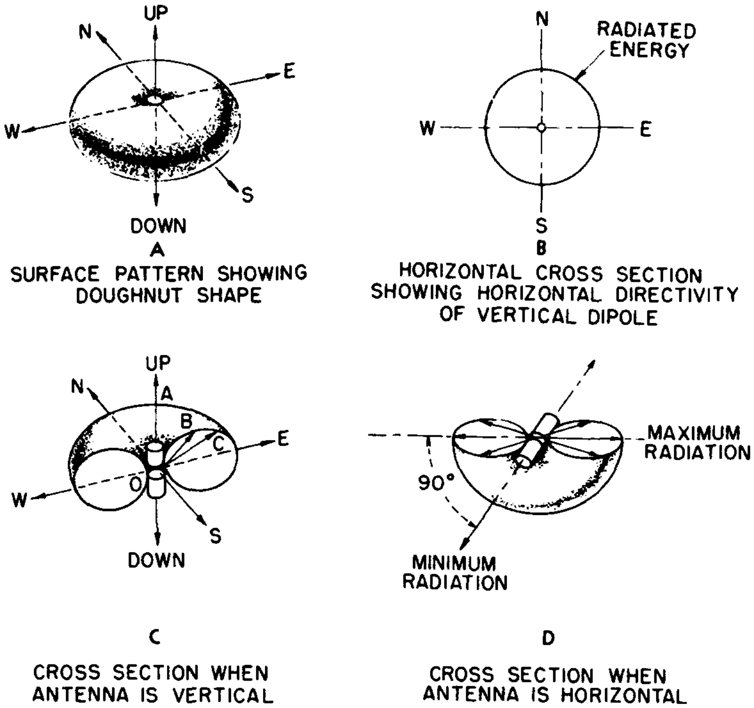


Figure 11-14.—Radiation pattern of a dipole.

tal cross-section pattern is shown in figure 11-14, B. Because a circular field pattern is created, the field strength is the same in any compass direction.

Theoretically a vertical dipole in free space has no vertical radiation along the direct line of its axis. However, it may produce a considerable amount of radiation at other angles measured to the line of the antenna axis. Figure 11-14, C, shows a vertical cross section of the radiation pattern of figure 11-14, A. The radiation along OA is zero; but at another angle, represented by angle *AOB*, there is appreci-

able radiation. At a greater angle, *AOC*, the radiation is still greater. Because of this variation in field strength pattern at different vertical angles, a field-strength pattern of a vertical half-wave antenna taken in a horizontal plane must specify the vertical angle of radiation for which the pattern applies.

Figure 11–14, D, shows half of the doughnut pattern for a horizontal half-wave dipole. The maximum radiation takes place in a plane perpendicular to the axis of the antenna and crossing through its center. A polar diagram representing the radiation pattern of a horizontal dipole is shown in figure 11–9.

ANTENNA COUPLING

A common method of coupling the shipboard communications antenna to its associated transmitter is shown in figure 11–15. The antenna tuning system is made up of the an-

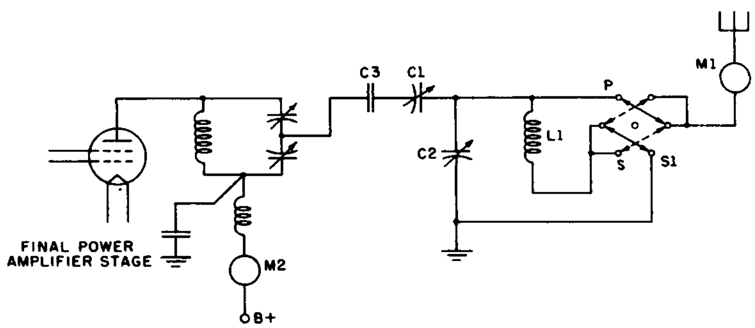


Figure 11–15.—Antenna coupling to transmitter.

tenna coupling capacitor, *C*1; the antenna tuning inductor, *L*1; the antenna tuning capacitor, *C*2; and the antenna feed switch, *S*1. The d-c blocking capacitor, *C*3, is connected in series with the antenna coupling capacitor, *C*1, to protect the antenna system from d-c potentials that might be occasioned by damage or voltage breakdown of the variable antenna coupling capacitor *C*1. The antenna tuning capacitor, *C*2, is variable, and is operated in one of two circuit arrange-

ments. With $S1$ in position P, capacitor $C2$ is connected in parallel with $L1$, and the antenna is voltage (or shunt) fed. With $S1$ in position S, capacitor $C2$ is connected in series with $L1$ and the antenna is current (or series) fed.

The antenna system is tuned by first adjusting $C1$ to minimum coupling and tuning the final power amplifier stage to resonance. Then capacitor $C2$ and inductor $L1$ are tuned for antenna resonance. The antenna now appears as a pure resistance to the final amplifier. The capacitance of $C1$ is then increased in small steps until the required loading of the final amplifier is obtained. However, each time the coupling capacitor is changed the final amplifier and antenna tuning circuit $L1$, $C2$ must be returned to resonance. Care must be taken not to overcouple with $C1$. After the final amplifier and the antenna circuit are resonated, the load on the final amplifier is purely resistive, and the maximum transfer of energy from the final amplifier to the antenna is obtained.

PROPAGATION OF RADIO WAVES

Radio Wave

When a radio wave leaves a vertical antenna the field pattern of the wave resembles a huge doughnut lying on the ground with the antenna in the hole at the center. Part of the wave moves outward in contact with the ground to form

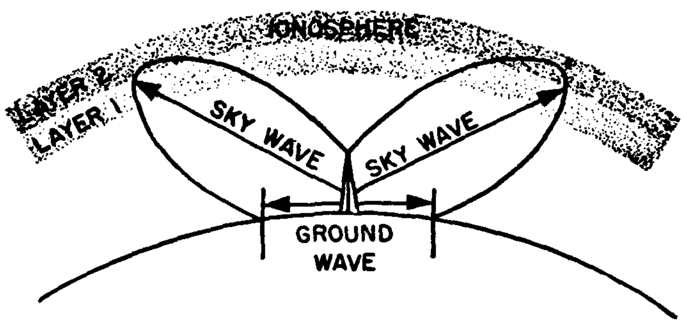


Figure 11–16.—Formation of the ground wave and sky wave.

the GROUND WAVE, and the rest of the wave moves upward and outward to form the SKY WAVE, as shown in figure 11–16. The ground and sky portions of the radio wave are responsible for two different methods of carrying the messages from transmitter to receiver. The ground wave is used both for short-range communication at high frequencies with low power, and for long-range communications at low frequencies with very high power. Daytime reception from most nearby commercial stations is carried by the ground wave.

The sky wave is used for long-range, high-frequency daylight communication. At night, the sky wave provides a means for long-range contacts at somewhat lower frequencies.

Ground Wave

The ground wave is commonly considered to be made up of two parts, a surface wave and a space wave. The surface wave travels along the earth's surface.

The space wave travels in the space immediately above the earth's surface in two paths—one directly from transmitter to receiver and the other a path in which the space wave is reflected from the ground before it reaches the receiver. Since the space wave follows two paths of different lengths, the two components may arrive in or out of phase with each other. Thus as the distance from the transmitter is changed, these two components may add or they may cancel. Neither of these component waves is affected by the reflecting layer of atmosphere high above the earth's surface called the IONOSPHERE.

The space-wave part of the ground wave becomes more important as the frequency is increased or as the transmitter and receiver antenna height is increased. When the transmitting and receiving antennas are both close to the ground, the space wave components cancel. This is true because the ground-reflected component is shifted 180° in phase upon reflection, has the same magnitude as the direct component, and travels a path of approximately the same length as that of the direct component. Thus the surface-wave part of the

ground wave is responsible for most of the daytime broadcast reception.

As it passes over the ground, the surface wave induces a voltage in the earth, setting up eddy currents. The energy to establish these currents is absorbed from the surface wave, thereby weakening it as it moves away from the transmitting antenna. Increasing the frequency, rapidly increases the attenuation so that surface-wave communication is limited to relatively low frequencies.

Shore-based transmitters are able to furnish long-range ground-wave communication by using frequencies between about 18 and 300 kc with extremely high power.

Since the electrical properties of the earth along which the surface wave travels are relatively constant, the signal strength from a given station at a given point is nearly constant. This holds true in nearly all localities except those that have distinct rainy and dry seasons. There the difference in the amount of moisture causes the soil's conductivity to change.

The conductivity of salt water is 5,000 times as great as that of dry soil. The superiority of surface-wave conductivity by salt water is the reason that high-power, low-frequency transmitters are located as close to the edge of the ocean as practicable.

Sky Wave

That part of the radio wave that moves upward and outward and that is not in contact with the ground is called the SKY WAVE. It behaves differently from the ground wave. Some of the energy of the sky wave is refracted (bent) by the ionosphere so that it comes back toward the earth. A receiver located in the vicinity of the returning sky wave will receive strong signals even though several hundred miles beyond the range of the ground wave.

Ionosphere

The ionosphere is found in the rarefied atmosphere approximately 40 to 350 miles above the earth. It differs from

the other atmosphere in that it contains a much higher number of positive and negative ions. The negative ions are believed to be free electrons. The ions are produced by the ultraviolet and particle radiations from the sun. The rotation of the earth on its axis, the annual course of the earth around the sun, and the development of sun spots all affect the number of ions present in the ionosphere, and these in turn affect the quality and distance of radio transmission.

The ionosphere is constantly changing. Some of the ions are recombining to form neutral atoms, while other atoms are being ionized by the removal of electrons from their outer orbits. The rate of formation and recombination of ions depends upon the amount of air present, and the strength of the sun's radiations.

At altitudes above 350 miles, the particles of air are too sparse to permit large-scale ion formation. Below about 40 miles altitude, only a few ions are present because the rate of recombination is too high. The sun's ultraviolet radiations have been absorbed in their passage through the upper layers of the ionosphere with the result that below an elevation of 40 miles too few ions exist to affect materially sky-wave communication.

Different densities of ionization at different heights make the ionosphere appear to have layers. Actually there is thought to be no sharp dividing line between layers, but for the purpose of discussion a sharp demarcation is indicated.

The ionized atmosphere at an altitude of between 40 and 50 miles is called the *D* LAYER. Its ionization is low and it has little effect on the propagation of radio waves except for the absorption of energy from the radio waves as they pass through it. The *D* layer is present only during the day. Its presence greatly reduces the field intensity of transmissions that must pass through daylight zones.

The band of atmosphere at altitudes between 50 and 90 miles contains the so-called *E* LAYER. It is a well-defined band with greatest density at an altitude of about 70 miles. This layer is strongest during the daylight hours, and is

also present but much weaker at night. The maximum density of the E layer appears at about noon local time.

The ionization of the E layer at the middle of the day is sometimes sufficiently intense to refract frequencies up to 20 megacycles back to the earth. This action is of great importance to daylight transmissions for distances up to 1,500 miles.

The F LAYER extends approximately from the 90-mile level to the upper limits of the ionosphere. At night only one F layer is present; but during the day, especially when the sun is high, this layer often separates into two parts, F_1 and F_2, as shown in figure 11–17. As a rule the F_2 layer is at its

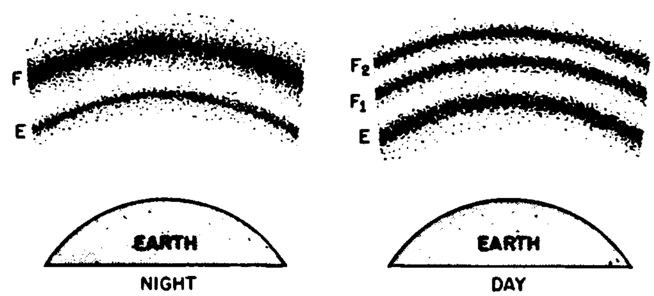


Figure 11–17.—E layer and F layer of the ionosphere.

greatest density during early afternoon hours, but there are many notable exceptions of maximum F_2 density existing several hours later. Shortly after sunset the F_1 and F_2 layers recombine into a single F layer.

In addition to the layers of ionized atmosphere that appear regularly, erratic patches of ionized atmosphere occur at E-layer heights in the manner that clouds appear in the sky. These patches are referred to as SPORADIC-E ionizations. They are often present in sufficient number and intensity to enable good v-h-f radio transmission over distances not normally possible.

Sometimes sporadic ionizations appear in considerable strength at varying altitudes and actually prove harmful to radio transmissions.

Effect of the Ionosphere on the Sky Wave

The ionosphere acts as a conductor, and absorbs energy in varying amounts from the radio wave. The ionosphere also acts as a radio mirror and refracts (bends) the sky wave back to the earth, as illustrated in figure 11–18.

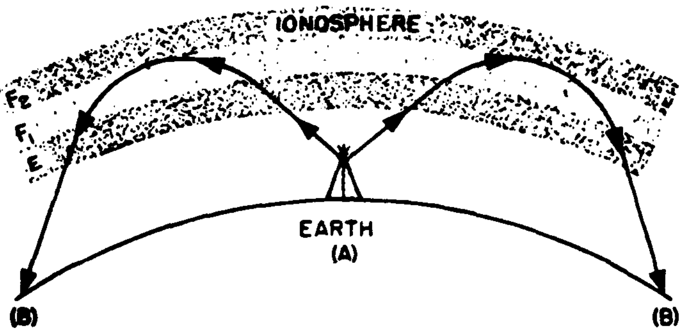


Figure 11–18.—Refraction of the sky waves by the ionosphere.

The ability of the ionosphere to return a radio wave to the earth depends upon the angle at which the sky wave strikes the ionosphere, the frequency of the transmission, and the ion density.

When the wave from an antenna strikes the ionosphere the wave begins to bend. If the frequency is correct (the ionosphere is suffiently dense and the angle is proper), the wave will eventually emerge from the ionosphere and return to the earth. If a receiver is located at either of the points *B* in figure 11–18, the transmission from point *A* will be received. The antenna height in the figure is not drawn to scale. The tallest antennas are not over 1,000 feet in height.

For example the sky wave in figure 11–19 is assumed to be composed of rays that emanate from the antenna in three distinct groups that are identified according to their angle of elevation. The angle at which the group-1 rays strike the ionosphere is too nearly vertical for the rays to be returned to the earth. The rays are bent out of line, but pass completely through the ionosphere and are lost.

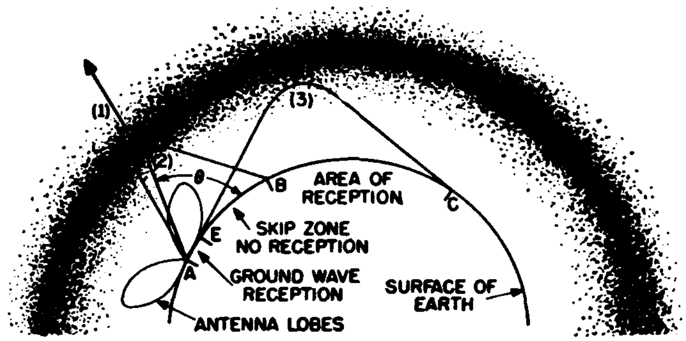


Figure 11–19.—Effect of the angle of departure on the area of reception

The angle made by the group-2 rays is called the CRITICAL ANGLE for that frequency. Any ray that leaves the antenna at an angle greater than this angle (θ) will penetrate the ionosphere.

Group-3 rays strike the ionosphere at the smallest angle that will be refracted and still return to the earth. At any smaller angle the rays will be refracted but will not return to the earth.

As the frequency increases, the critical angle decreases. Low-frequency fields can be projected straight upward and will be returned to the earth. The highest frequency that can be sent directly upward and still be returned to the earth is called the CRITICAL FREQUENCY. At sufficiently high frequencies, regardless of the angle at which the rays strike the ionosphere, the wave will not be returned to the earth. The critical frequency is not constant but varies from one locality to another, with the time of day, with the season of the year, and with the sunspot cycle.

Because of this variation in the critical frequency, nomograms and frequency tables are issued that predict the maximum usable frequency (MUF) for every hour of the day for every locality in which transmissions are made.

Nomograms and frequency tables are prepared from data obtained experimentally from stations scattered all over the world. All this information is pooled, and the results are

tabulated in the form of long-range predictions that remove most of the guess work from radio communication.

In the example in figure 11–19 the area between points *B* and *C* will receive the transmission by way of the refracted sky wave. The area between points *A* and *E* will receive the transmission by ground wave. All receivers located in the SKIP ZONE between points *E* and *B* will receive no transmissions from point *A*, because neither the sky wave nor the ground wave reaches this area.

Effect of Daylight on Wave Propagation

The increased ionization during the day is responsible for several important changes in sky-wave transmission. It causes the sky wave to be returned to the earth nearer to the point of transmission. The extra ionization increases the absorption of energy from the sky wave; if the wave travels a sufficient distance into the ionosphere, it will lose all of its energy. The presence of the F_1 and E layers with the F_2 layer make long-range, high-frequency communications possible, provided the correct frequencies are used.

Absorption usually reduces the effective daylight communication range of low-frequency and medium-frequency transmitters to surface-wave ranges.

High-Frequency Long-Range Communications

The high degree of ionization of the F_2 layer during the day, enabling refraction of high frequencies which are not greatly absorbed, has an important effect on transmissions of the high-frequency band. Figure 11–20 shows how the F_2

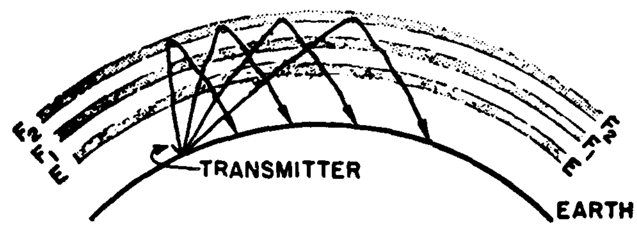


Figure 11–20.—*Effect of the F_2 layer on transmission of high-frequency signals.*

layer completes the refraction and returns the transmissions of these frequencies to the earth, thereby making possible long-range high-frequency communication during the daylight hours.

The waves are partially bent in going through the E layer and the F_1 layer, but are not returned to the earth until the F_2 layer completes the refraction process. V-h-f waves pass directly through the ionosphere.

The exact frequency to be used to communicate with another station depends upon the condition of the ionosphere and the distance between stations. Since the ionosphere is constantly changing, the nomograms and frequency tables are used to select the correct frequency for the desired distance and for the time of day the transmission is to be made.

Multiple Refraction

The radio wave may be refracted many times between the transmitter and receiver locations, as shown in figure

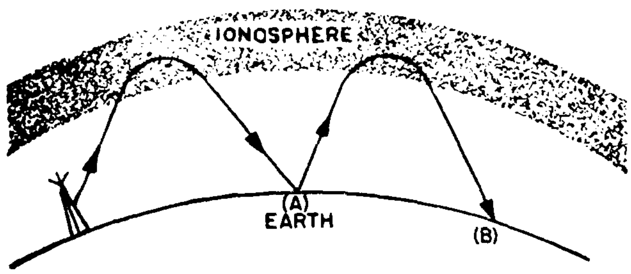


Figure 11-21.—Multiple refraction and reflection of a sky wave.

11-21. In this example the radio wave strikes the earth at location A, with sufficient intensity to be reflected back to the ionosphere and there to be refracted back to the earth a second time. Frequently a sky wave has sufficient energy to be refracted and reflected several times, thus greatly increasing the range of transmission. Because of this so-

called MULTIPLE-HOP transmission, transoceanic and around-the-world transmission is possible with moderate power.

Fading

Fading is a term used to describe the variations in signal strength that occur at the receiver during the time a signal is being received. There are several reasons for fading; some are easily understood while others are more complicated.

One cause is probably the direct result of interference between single-hop (fig. 11–18) and double-hop transmissions (fig. 11–21) occurring simultaneously from the same source. If the two waves arrive in phase, the signal strength will be increased, but if the two waves arrive in phase opposition (180° out of phase), they will cancel each other, and the signal will be weakened.

Interference fading also occurs where the ground wave and sky wave come in contact with each other. This type of fading becomes severe if the two waves are approximately equal in strength. Fluctuations in the sky wave with a steady ground wave can cause worse fading than sky-wave transmission alone.

Variations in absorption and in the length of the path in the ionosphere are also responsible for fading. Occasionally, sudden disturbances in the ionosphere cause complete absorption of all sky-wave radiations.

Receivers located near the outer edge of the skip zone are subjected to fading as the sky wave alternately strikes and skips over the area. This type of fading sometimes causes the received signal strength to fall to nearly the zero level.

Frequency Blackouts

Frequency blackouts are closely related to certain types of fading, some of which are severe enough to completely blank out the transmission.

Changing conditions in the ionosphere shortly before sunrise and shortly after sunset may cause complete blackouts at certain frequencies. The higher frequency signals pass

through the ionosphere while the lower frequency signals are absorbed by it.

Ionospheric storms (turbulent conditions in the ionosphere) often cause radio communication to become erratic. Some frequencies will be completely blacked out, while others may be reinforced. Sometimes these storms develop in a few minutes, and at other times they require as much as several hours to develop. A storm may last several days.

When frequency blackouts occur, the operator must be alert if he is not to lose contact with other ship or shore stations. In severe storms the critical frequencies are much lower, and the absorption in the lower layers of the ionosphere is much greater.

V-H-F and U-H-F Communication

In recent years there has been a trend toward the use of frequencies above 30 megacycles for short-range ship-to-ship and ship-to-airplane communications.

Early concepts suggested that these transmissions traveled in straight lines. This leads to the assumption that the u-h-f transmitter and receiver must be within sight of each other in order to supply radio contact.

Extensive use and additional research show the early "line-of-sight" theory to be frequently in error because radio waves of these frequencies may be refracted. The receiver does not always have to be in sight of the transmitter. Although this type of transmission still is called LINE-OF-SIGHT transmission, it is better to call it v-h-f and u-h-f transmission.

In general the v-h-f and u-h-f waves follow approximately straight lines, and large hills or mountains cast a radio shadow over these areas in the same way that they cast a shadow in the presence of light rays. A receiver located in a radio shadow will receive a weakened signal and in some cases, no signal at all. Theoretically, the range of contact is the distance to the horizon, and this distance is determined by the heights of the two antennas. However, as stated previously, communication is sometimes possible many

hundreds of miles beyond the assumed horizon range. This fact must be observed when transmission is to occur under conditions of radio security.

Effect of Atmosphere on High-Frequency Transmissions

Unusual ranges of v-h-f and u-h-f contacts are caused by abnormal atmospheric conditions a few miles above the earth. Normally the warmest air is found near the surface of the water. The air gradually becomes cooler as the altitude increases. Sometimes unusual situations develop where warm layers of air are found above cooler layers. This condition is known as temperature inversion.

When a temperature inversion exists, the amount of refraction (index of refraction) is different for the particles trapped within the boundaries than it is for those outside them. These differences form channels or ducts that will conduct the radio waves many miles beyond the assumed normal range.

Sometimes these ducts are in contact with the water and may extend a few hundred feet into the air. At other times the duct will start at an elevation of between 500 and 1,000 feet and extend an additional 500 to 1,000 feet in the air.

If an antenna extends into the duct or if the wave enters a duct after leaving an antenna, the transmission may be conducted a long distance. An example of this type of transmission of radio waves in ducts formed by temperature inversions is shown in figure 11-22.

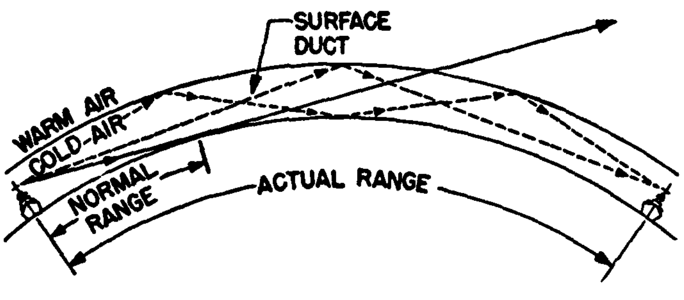


Figure 11-22.—Duct effect in high-frequency transmission.

With certain exceptions ducts are formed over water where the following conditions are observed aboard ship:

1. A wind is blowing from land.
2. There is a stratum of quiet air.
3. There are clear skies, little wind, and high barometric conditions.
4. A cool breeze is blowing over warm open ocean, especially in the tropic areas and in the trade-wind belt.
5. Smoke, haze, or dust fails to rise, but spreads out horizontally.
6. The moisture content of the air at the bridge is considerably less than at the sea's surface.
7. The temperature at the bridge is a few degrees higher than at the surface of the water.
8. The received signal is fading rapidly.

General Use of Frequencies

Each frequency band has its own special uses. These uses depend upon the nature of the waves—surface, sky, or space, and the effect that the sun, the earth, the ionosphere, and the earth's atmosphere have upon them.

It is difficult to establish fixed rules for the choice of a frequency for a particular purpose. Some general statements can be made however as to which frequency bands are best suited to the type of transmission to be made.

For example, if a long-range communication is to get through to a distant receiver, high power and low frequency should be used. The large international communication systems and the large FOX stations use this combination of high power and low frequency. However, this combination requires an antenna array that may be too large for use with shipboard transmission. An alternative is to transmit the message to the nearest shore station for relay to its destination.

The sky wave builds up to a peak of usefulness in the h-f band. At night the peak of daytime usefulness is in the top third of the m-f band. The usefulness of the ground, or

surface, wave declines steadily as the higher frequencies are reached until it is of no value in the h-f band. The only means of radio communication in the v-l-f band and for a limited range above the v-l-f band is the space wave component of the ground wave.

Sky-wave transmission (1,600 to 30,000 kc) is almost always associated with skip distances. Great range can be obtained, but in the process many receiving stations may be skipped in between the source and the most remote point where the signal is to be received. Thus one of the receiving stations to which it is most desired to get the signal may be skipped if the receiving station is located within any one of the skip zones associated with the frequency being transmitted.

The most important frequencies for long-range transmission are those from 2,000 to 18,000 kc (2 to 18.1 mc). This band is standard for long-distance naval communications from ship to ship and ship to shore. It is the band that is used most frequently and the one that is covered by the standard Navy transmitters such as the TBK, TBL, and TBM. The band is in the short-wave region, and transmissions are accomplished by means of the sky wave and are affected by skip distances. When long range is desired in daytime the frequencies that should be used are approximately from 7 megacycles to 18 megacycles. For night communications, frequencies below about 10 or 15 megacycles should be used.

Nonregistered Publications Memoranda (NRPM's), which are supplied to the various ships of the Navy, contain tables that show the best frequencies within the band for communication with various shore stations. These tables give the recommended frequency for every hour of the day for distances varying from 250 to 5,000 miles for some stations. The direction of the receiving station from the ship transmitting the signals is also taken into account. NRPM's cover a 3-month period with a separate table for each month and for each major shore station.

QUIZ

1. State a basic principle of the radiation of electromagnetic energy that pertains to a moving electric field and its associated magnetic component; also to a moving magnetic field and its associated electric field component.
2. State the formula for the wavelength of an electromagnetic wave in free space in terms of the frequency of the radiating source and the speed of the wave. (Assume the wavelength is in meters and the frequency is in megacycles.)
3. If the r-f generator supplying a tuned dipole antenna has a voltage output of sine waveform, what is the waveform of the antenna current distribution with respect to the length of the dipole?
4. At what approximate distance in wavelengths from an antenna does the induction field become negligible?
5. How does the strength of the radiated field vary with distance?
6. Why are resonant conductors more efficient radiators than nonresonant conductors?
7. What is the relative velocity of the radio wave in an antenna compared to that of the wave in free space?
8. What determines the magnitude of the antenna current at the feedpoint for a given r-f voltage at that point?
9. Why is a large-diameter radiator physically shorter than a small-diameter radiator, assuming the same resonant frequency in both cases?
10. Define radiation resistance.
11. What is the approximate radiation resistance of e[illegible]e following current-fed antennas?
    a. Half-wave
    b. Quarter-wave
12. What are some of the smaller losses that are found in an antenna system?
13. What is the relative position of a simple radiating antenna with respect to the electric lines of the radiated wave?
14. When the transmitting antenna is several wavelengths above ground, which type of polarization gives the stronger signal close to the earth?
15. What are the distinguishing characteristics of a Hertz antenna?
16. What are the distinguishing characteristics of a Marconi antenna?
17. How may an antenna of fixed length be "lengthened" electrically?
18. How may an antenna be "shortened" electrically?

19. In figure 11–15, what type of impedance (R, X_L, or X_C) is offered to the final amplifier tank when the coupling circuits are properly adjusted?
20. When the transmitting and receiving antennas are close to the ground, why will the direct component and the ground-reflected component of the space wave cancel?
21. Why is surface-wave reception limited to relatively low frequencies?
22. What is the relative number of positive and negative ions that exist in the ionosphere compared with the number contained in other portions of the earth's atmosphere?
23. What is the layer designation of that portion of the ionized atmosphere having an altitude of between 40 and 50 miles that is present only during the day?
24. Upon what three factors does the ability of the ionosphere to return a radio wave to the earth depend?
25. As the frequency increases, what happens to the critical angle made by the group-2 rays of figure 11–19?
26. What is the effect on the range of low- and medium-frequency daylight communications, of increased absorption of energy from the sky wave with increase in ionization?
27. Which layer of the ionosphere makes possible long-range high-frequency communications during daylight hours?
28. Are v-h-f and u-h-f always limited to line-of-sight transmission?
29. Describe the atmospheric condition known as temperature inversion.
30. What is the effect of a temperature inversion a few miles above the earth on the range of v-h-f and u-h-f transmissions?

CHAPTER 12

ELEMENTARY COMMUNICATIONS RECEIVERS

INTRODUCTION

Many of the principles, circuits, and components discussed in previous chapters are directly applicable to radio receivers. A basic knowledge of them is therefore assumed in treating the material included in this chapter.

At the radio transmitter the carrier frequency is modulated by the desired signal, which may consist of coded characters, voice, music, or other types of signals. AMPLITUDE MODULATION (a-m) occurs if the signals cause the amplitude of the carrier to vary. FREQUENCY MODULATION (f-m) occurs if the signals cause the frequency of the carrier, or center frequency, to vary. Although there are other types of modulation, only a-m and f-m will be treated in this chapter.

The r-f carrier wave with the modulating signal impressed upon it is transmitted through space as an electromagnetic wave to the antenna of the receiver. As the wave passes across the receiving antenna, small a-c voltages are induced in the antenna. These voltages are coupled into the receiver via the antenna coupling coil. The function of the receiver is to select the desired carrier frequency from those present in the antenna circuit and to amplify the small a-c signal voltage. The receiver then removes the carrier by the process of detection (rectification and the removal of the r-f

component) and amplifies the resultant audio signal to the proper magnitude to operate the loudspeaker or earphones.

Three major types of radio receivers are reviewed in this chapter—the TUNED-RADIO-FREQUENCY (t-r-f) receiver, the SUPERHETERODYNE receiver, and the FREQUENCY-MODULATION (f-m) receiver. The a-m and f-m superheterodyne receivers are often combined in one set, because many of the circuit components are common to both types of receivers.

T-R-F RECEIVERS

Operating Principle

The tuned-radio-frequency receiver, generally known as the t-r-f receiver, consists of one or more r-f stages, a detector stage, one or more a-f stages, a reproducer, and the necessary power supply. A block diagram of a t-r-f receiver is shown in figure 12-1. The waveforms that appear in the

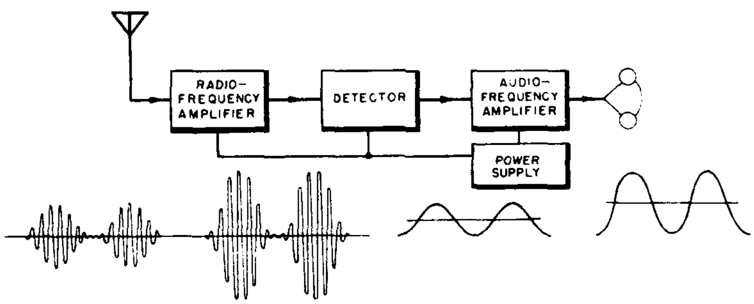


Figure 12-1.—Block diagram of a t-r-f receiver and waveforms.

respective sections of the receiver are shown below the block diagram.

The amplitude of the a-m signal at the input of the receiver is relatively small because it has been attenuated in the space between the transmitter and the receiver. It is composed of the carrier frequency and the modulation envelope. The r-f amplifier stages amplify the waveform, but they do not change its basic shape if the circuits are operating properly.

The detector rectifies and removes the r-f component of the signal. The output of the detector is a weak signal made up only of the modulation component, or envelope, of the incoming signal. The a-f amplifier stages following the detector increase the amplitude of the a-f signal to a value sufficient to operate the loudspeaker or earphones.

Components

R-F SECTION.—The ANTENNA-GROUND SYSTEM serves to introduce the desired signal into the first r-f amplifier stage via the antenna coupler transformer. For best reception the resistance of the antenna-ground system should be low. The aerial should also be of the proper length for the band of frequencies to be received; and the antenna impedance should match the input impedance of the receiver. The gain of most commercial receivers, however, is generally sufficient to make these values noncritical.

The R-F AMPLIFIERS in the t-r-f receiver have tunable tanks in the grid circuits. Thus, the receiver may be tuned so that only one r-f signal within its tuning range is selected for amplification. When the tank is tuned to the desired frequency, it resonates and produces a relatively large circulating current. The grid of the r-f amplifier then receives a relatively large signal voltage at the resonant frequency, and minimum signals at other frequencies.

The relative ability of a receiver to select one particular frequency and to reject all others is called the SELECTIVITY of the receiver. The relative ability of the receiver to amplify small signal voltages is called the SENSITIVITY of the receiver. Both of these values may be improved by increasing the number of r-f stages. When this is done, the tuning capacitors in the grid tank circuits are usually ganged on the same shaft and trimmers are added in parallel with each capacitor to make the stages track at the same frequency. In addition, the outer plates of the rotor sections of the capacitors are sometimes slotted to enable more precise alignment throughout the tuning range.

Tetrodes or pentodes are generally used in r-f amplifiers

because, unlike triodes, they do not usually require neutralization. They also have higher gain than triodes.

A typical r-f amplifier stage employing a pentode is shown in figure 12-2. Tuned circuit $L2C1$ is inductively coupled

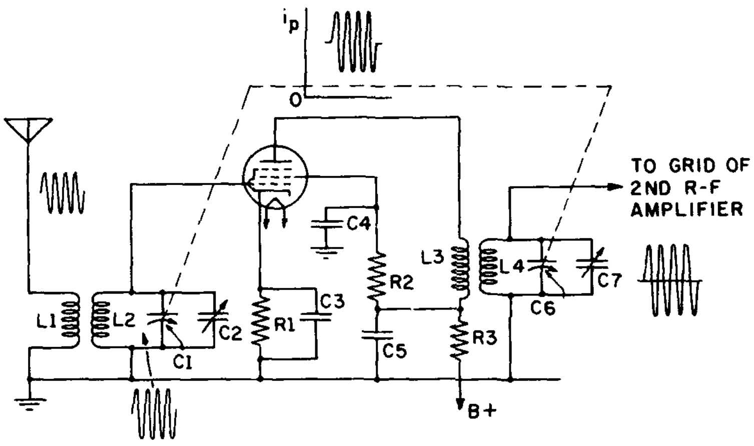


Figure 12-2.—R-f amplifier stage.

to $L1$, the antenna coil. $R1$ and $C3$ provide a smooth operating bias for the tube. $C4$ and $R2$ are the screen bypass capacitor and dropping resistor, respectively. The tuned circuit, L4C6, couples the following stage inductively to $L3$. Both transformers are of the air-core type. The dotted lines indicate mechanical ganging of $C1$ and $C6$ on the same shaft. The tuning capacitor in the next stage is also ganged on the same shaft.

If it is desired to cover more than one frequency range, additional coils having the proper inductance are used. They are sometimes of the plug-in variety, but more generally they are mounted on the receiver and their leads are connected to a multicontact rotary switch. The latter method is preferable for BAND SWITCHING because the desired band can be selected simply by turning the switch. The same tuning capacitor is used for each band. However, when band switching is employed, the trimmers are connected

across the individual tuning inductors and not across the main tuning capacitors.

When two or more amplifier stages are supplied from a common B supply, feedback occurs as a result of common coupling between the plate circuits, and some method of decoupling must be employed. The coupling consists of the internal impedance of the source of plate voltage. The feedback may either increase or decrease the amplification depending on the phase relation between the input voltage and the feedback voltage. In a multistage amplifier the greatest transfer of feedback energy occurs between the final and first stages because of the high amplification through the multistage amplifier.

The effects of feedback are important if the feedback voltage coupled into the plate circuit of the first stage is appreciable compared to the signal voltage that would be developed if feedback did not exist. For example, a three-stage resistance-coupled amplifier may develop a feedback voltage (coupled via the B supply into the plate circuit of stage 1) which is in phase with the signal voltage of stage 1 and hence may cause oscillations to be set up. In audio amplifiers having high gain and a good low-frequency response this regeneration causes a low-frequency oscillation known as "motorboating" because of the "putt-putt" sound in the speaker.

Decoupling circuits are designed for both r-f and a-f amplifiers to counteract feedback. Thus, in the r-f amplifier in figure 12-2, *C*5 and *R*3 make up the decoupling circuit. *R*3 offers a high impedance to the signal current, but *C*5 offers a low impedance. Consequently the signal current is shunted to ground around the B supply. Since *R*3 also offers a high resistance to d-c current, it may be replaced by a choke coil having a high impedance only to the signal current. Each stage is similarly equipped with a decoupling circuit.

A mechanical or an electrical bandspread may be used as an aid in separating stations that are crowded together on the tuning dial. MECHANICAL BANDSPREAD is simply a

micrometer arrangement to reduce the motion of the capacitor rotor as the tuning knob is turned. When ELECTRICAL BANDSPREAD is used a small variable capacitor is connected in parallel with the tuning capacitor. Because of its small size, this variable capacitor may be moved a considerable amount before it causes an appreciable change in the frequency of the tuned circuit. If the tuning capacitors are ganged, the bandspread capacitors are also ganged.

DETECTOR.—The process of removing the intelligence component of the modulated wave from the r-f carrier is called DETECTION OR DEMODULATION. In the a-m system the audio or intelligence component causes both the positive and the negative half cycles of the r-f wave to vary in amplitude. The function of the detector is to rectify the modulated signal. A suitable filter eliminates the remaining r-f pulses and passes the audio component on to the a-f amplifiers.

Details of the various methods of detection are treated in chapter 8. Each of the several methods that might be used in the t-r-f receiver have certain inherent weaknesses. For example, the diode detector requires several stages of amplification ahead of the detector. It loads its tuned input circuit, and therefore the sensitivity and selectivity of the circuit are reduced. However, it can handle strong signals without overloading, and its linearity is good.

The grid-leak detector is sensitive (and therefore requires fewer stages of amplification), but it has poor linearity and selectivity and it may be overloaded on strong signals.

The regenerative detector is very sensitive and has excellent selectivity, but it has poor linearity and easily overloads on strong signals.

The circuit shown in figure 12–3 employs plate detection. It has medium sensitivity and the ability to handle strong signals without overloading. The selectivity of this circuit is excellent, but because the i_p-e_g graph is curved near the cutoff point (where the plate detector operates) some distortion in the output cannot be avoided.

In figure 12–3, the tube is biased nearly to cutoff by the average plate current that flows through $R1$. This average

value increases as the signal strength increases. On positive half cycles of the incoming signals the plate current varies with the amplitude of the modulating wave and produces the desired a-f output voltage. On negative half cycles no

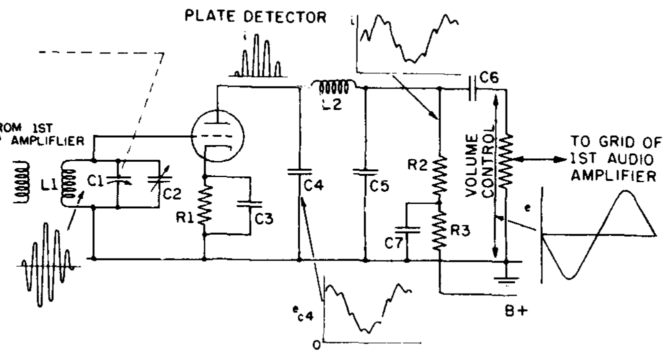


Figure 12-3.—Plate detector circuit.

appreciable plate current flows. Between positive half cycles the bias voltage is held constant across *R*1 by the action of *C*3, because the time constant of *R*1*C*3 is long compared with the time for the lowest a-f cycle.

The r-f pulses are filtered out by means of a low-pass filter (consisting of *C*4, *L*2, and *C*5), which rejects the r-f component and passes the a-f component. *C*6 couples the a-f component to the first audio amplifier. *R*2 is the plate load resistor, and the combination *R*3*C*7 makes up the decoupling circuit.

A–F SECTION.—The function of the a-f section of a receiver is to further amplify the audio signal, which is commonly fed via the volume control to the grid of the first audio amplifier tube. In most cases the amount of amplification that is necessary depends on the type of reproducer used. If the reproducer consists of earphones, only one stage of amplification may be necessary. If the reproducer is a large speaker or other mechanical device requiring a large amount of

power, several stages may be necessary. In most receivers the last a-f stage is operated as a power amplifier.

A necessary part of the a-f section is some means of manual control of the output signal level of the receiver.

A MANUAL VOLUME CONTROL may be employed in a number of receiver circuits. Normally this control varies the amplitude of the signal applied to the grid of an amplifier tube, as shown in figure 12–3. Increasing the resistance between ground and the sliding contact increases the amplitude of the signal applied to the grid of the driven stage.

An A-F OUTPUT STAGE is shown in figure 12–4. *C*1 couples

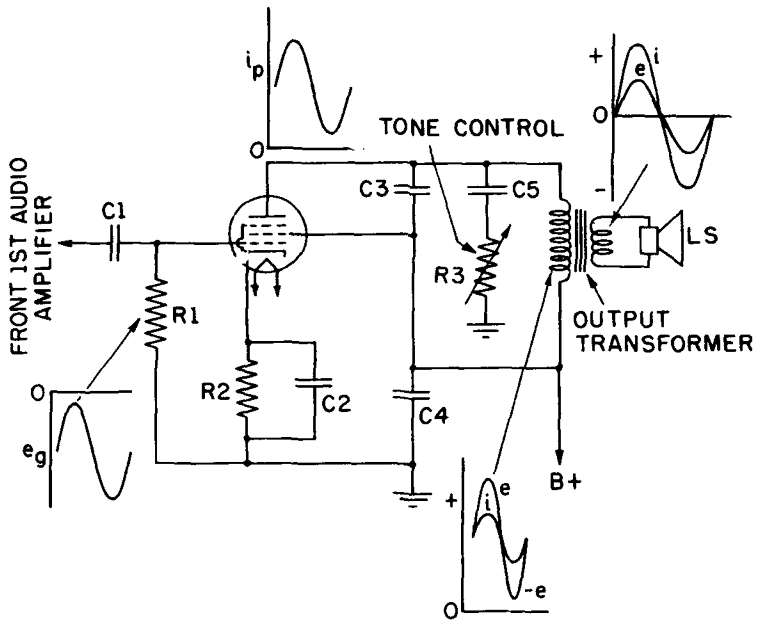


Figure 12–4.—Audio amplifier output stage.

the first a-f amplifier to the output stage, and *R*1 is the grid coupling resistor. *R*2 and *C*2 provide a steady bias. Because of the low frequencies involved, *C*2 should have a larger value of capacitance than similar bypass capacitors

in the r-f section. $C4$ is the plate-bypass, or decoupling, capacitor. $C3$ has a small value of capacitance and bypasses some of the higher frequencies around the output transformer, thus emphasizing the bass. The impedance of the output-transformer primary commonly represents a compromise between maximum power transfer and minimum distortion. The impedance of the secondary is chosen to match the impedance of the voice coil. Some secondaries have taps on the windings to permit an impedance match to a variety of voice-coil impedances.

TONE CONTROL may be used in communications receivers. The purpose of tone control is to emphasize either the low or the high frequencies, by shunting the undesired frequencies around the remainder of the circuit components in the audio section. A simple tone-control circuit, such as the series capacitor $C5$ and variable resistor $R3$ combination shown in figure 12–4, may be connected between plate and ground or between grid and ground in any of the audio stages of a receiver. In this figure it is connected between plate and ground. The value of the series capacitor is such that it will bypass to ground the high-frequency components. The amount of high-frequency energy removed by the tone-control circuit is determined by the setting of the variable-resistor control arm. When the resistance is low, the high frequencies are attenuated; when it is high they appear in the output.

LOUDSPEAKERS AND EARPHONES.—Figure 12–5 shows three types of audio reproducers.

In the permanent-magnet dynamic type of reproducer (fig. 12–5, A) a strong field is established between the pole pieces by means of a powerful permanent magnet. The flux is concentrated in the air gap between a permeable soft-iron core and an external yoke. The voice coil is mounted in the air gap. When a-c signal currents flow in the coil a force proportional to the strength of the current is applied to the coil, and the coil is moved axially in accordance with the a-c signal. The loudspeaker diaphragm is attached to the voice coil and moves in accordance with the

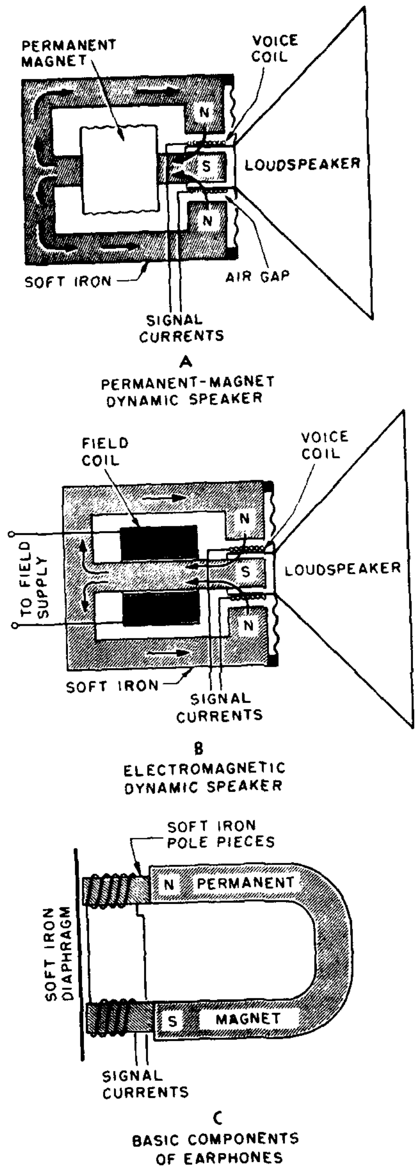


Figure 12-5.—Types of audio reproducers.

signal currents, thus setting up sound waves in the air. The corrugated diaphragm to which the speaker cone is attached keeps the cone in place and properly centered.

As in figure 12–5, B, an electromagnet may be used in place of the permanent magnet to form an electromagnetic dynamic speaker. However, in this instance sufficient d-c power must be available to energize the field. The operation is otherwise much the same as that of the permanent-magnet type.

The basic components of earphones are shown in figure 12–5, C. When no signal currents are present, the permanent magnet exerts a steady pull on the soft-iron diaphragm. Signal current flowing through the coils mounted on the soft-iron pole pieces develops a magnetomotive force that either adds to or subtracts from the field of the permanent magnet. The diaphragm thus moves in or out according to the resultant field. Sound waves will then be reproduced that have amplitude and frequency (within the capability of the reproducer) similar to the amplitude and frequency of the signal currents.

Circuit of the T-R-F Receiver

The complete circuit of a t-r-f radio receiver operated from an a-c power supply is shown in figure 12–6. The receiver uses two pentodes in the r-f section, one triode operated as a plate detector, and two pentode a-f amplifier stages that feed the loudspeaker.

From previous discussions, the various circuits may be identified and the signal may be traced from the antenna-ground system to the loudspeaker. The dotted lines indicate that the three main tuning capacitors are ganged on a single shaft. Across each of the main tuning capacitors is connected a trimmer capacitor to enable circuit alignment. The ground circuit and the various decoupling circuits may be readily identified. The power supply voltage is obtained from a conventional full-wave rectifier. Rectifier and tube filament currents are obtained from two low-voltage windings on the power transformer.

Characteristics of the T-R-F Receiver

The principal disadvantage of the t-r-f receiver is that its selectivity, or its ability to separate signals, does not remain constant over its tuning range. As the set is tuned from the low-frequency end of its tuning range to the high-frequency end, its selectivity decreases.

Also, the amplification, or gain, of a t-r-f receiver is not constant over the tuning range of the receiver. The gain depends on r-f transformer gain, which increases with frequency. In order to improve the gain at the low-frequency end of the band, r-f transformers employing high-impedance (untuned) primaries are designed so that the primary inductance will resonate with the primary distributed capacitance at some frequency slightly below the low end of the tunable band. Thus, the gain is good at the low end of the band because of the resonant build up of primary current. The near-resonant condition of the primary at the low end more than offsets the effect of reduced transformer action. However, the shunting action of the primary distributed capacitance lowers the gain at the high-frequency end of the band. To make up for the resultant poor gain at the high end of the band, a small capacitor is connected between the plate and grid leads of adjacent r-f stages to supplement the transformer coupling. At the low end of the band the capacitive coupling is negligible.

The superheterodyne receiver has been developed to overcome many of the disadvantages of the t-r-f receiver.

SUPERHETERODYNE RECEIVERS

Operating Principle

The essential difference between the t-r-f receiver and the superheterodyne receiver is that in the former the r-f amplifiers preceding the detector are tunable over a band of frequencies; whereas in the latter the corresponding amplifiers are tuned to one fixed frequency called the INTERMEDIATE FREQUENCY (i-f). The principle of frequency conversion by heterodyne action is here employed to reduce any desired

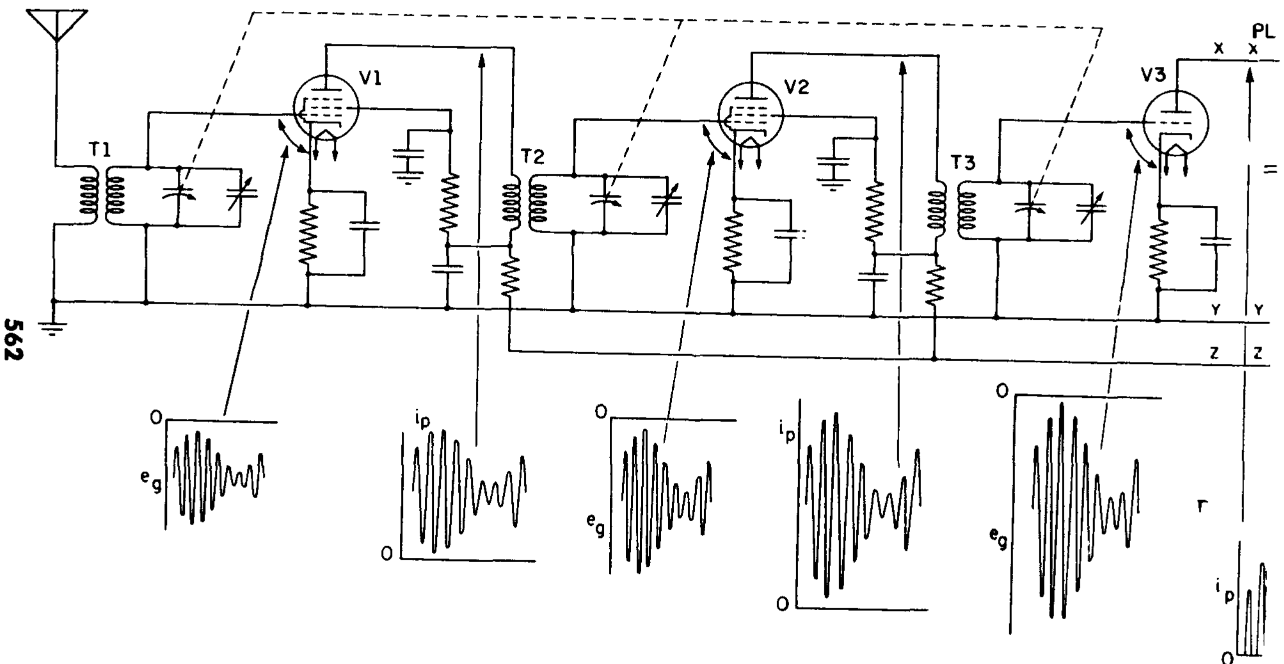


Figure 12-6.—Circuit of a t-r-f receiver—Continued

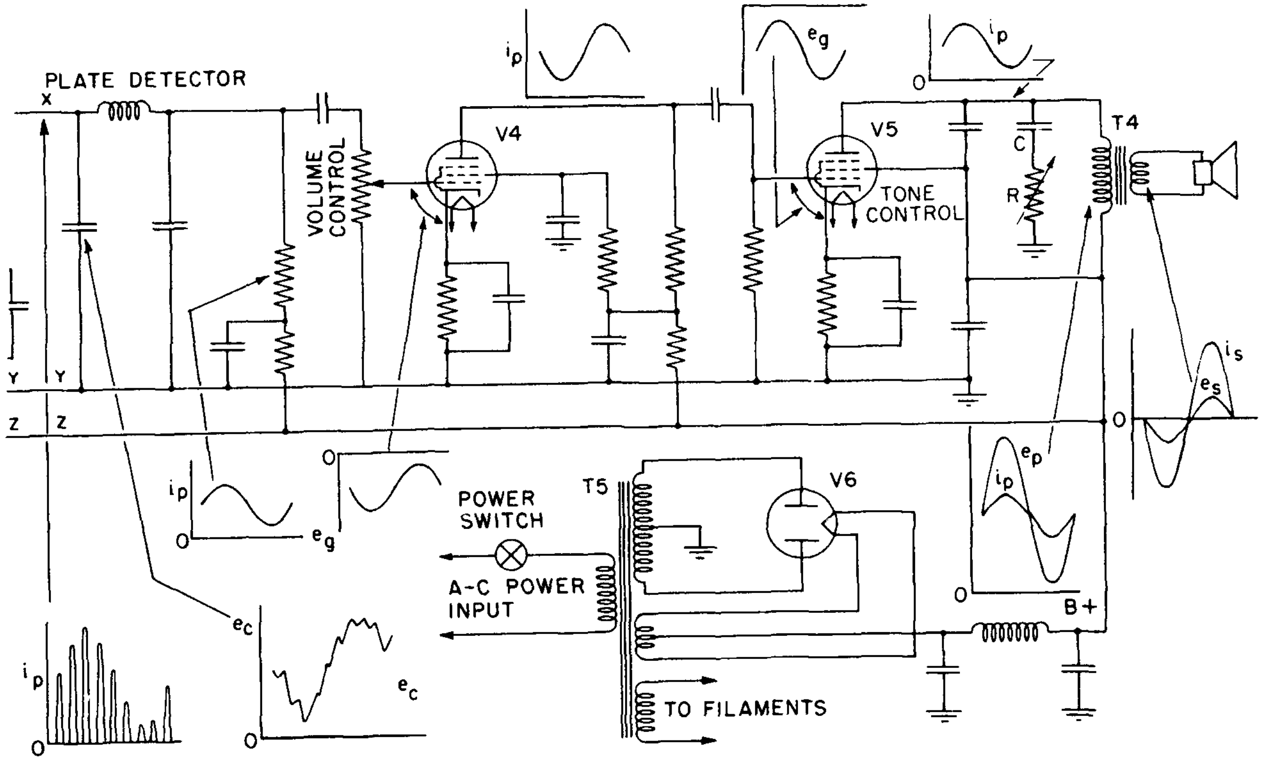


Figure 12-6.—Circuit of a t-r-f receiver.

station frequency within the receiver range to this intermediate frequency. Thus an incoming signal is converted to the fixed intermediate frequency before detecting the audio signal component, and the i-f amplifier operates under uniformly optimum conditions throughout the receiver range. The i-f circuits thus may be made uniformly selective, uniformly high in voltage gain, and uniformly of satisfactory bandwidth to contain all of the desired side-band components associated with the amplitude-modulated carrier.

The block diagram of a typical superheterodyne receiver is shown in figure 12–7. Below corresponding sections of the receiver are shown the waveforms of the signal at that point. The r-f signal from the antenna passes first through an r-f amplifier (preselector) where the amplitude of the signal is increased. A locally generated unmodulated r-f signal of constant amplitude is then mixed with the carrier frequency in the mixer stage. The mixing or heterodyning of these two frequencies produces an intermediate-frequency signal which contains all of the modulation characteristics of the original signal. The intermediate frequency is equal to the difference between the station frequency and the oscillator frequency associated with the heterodyne mixer. The intermediate frequency is then amplified in one or more stages called INTERMEDIATE-FREQUENCY (i-f) AMPLIFIERS and fed to a conventional detector for recovery of the audio signal.

The detected signal is amplified in the a-f section and then fed to a headset or loudspeaker. The detector, the a-f section, and the reproducer of a superheterodyne receiver are basically the same as those in a t-r-f set, except that diode detection is generally used in the superheterodyne receiver. Automatic volume control or automatic gain control also is commonly employed in the superheterodyne receiver

R-F AMPLIFIER.—If an r-f amplifier is used ahead of the mixer stage of a superheterodyne receiver it is generally of conventional design. Besides amplifying the r-f signal, the

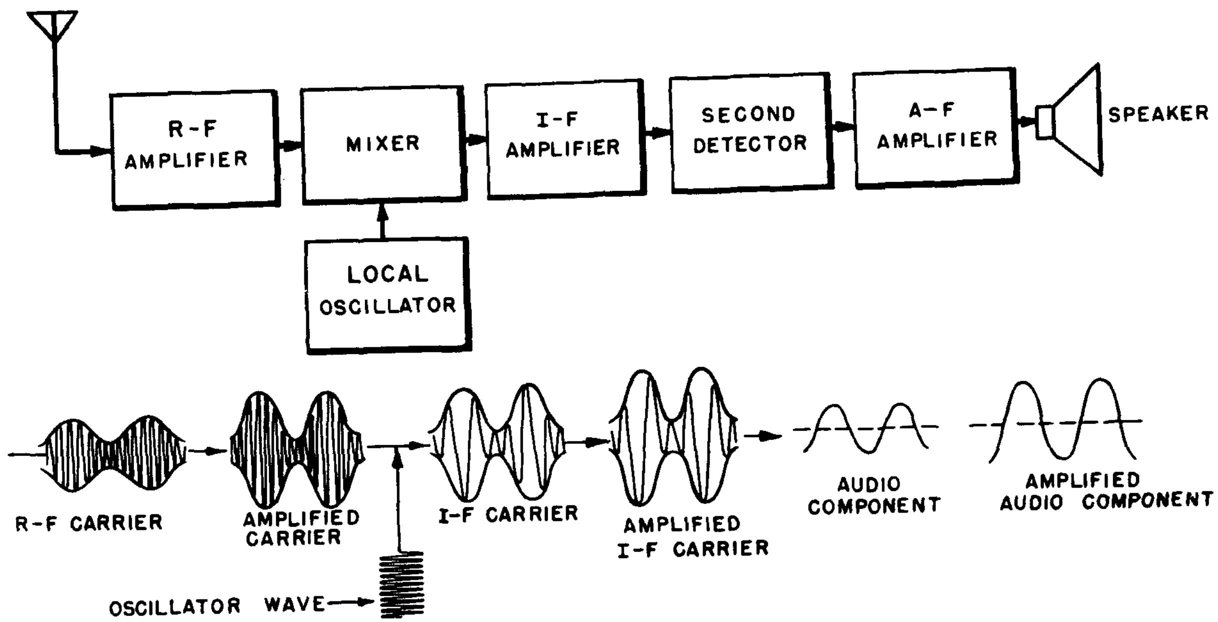


Figure 12-7.—Block diagram of a superheterodyne receiver and waveforms.

r-f amplifier has other important functions. For example, it isolates the local oscillator from the antenna-ground system. If the antenna were connected directly to the mixer stage, a part of the local oscillator signal might be radiated into space. This signal could be picked up by a sensitive direction finder on any enemy ship. For this reason and others, Navy superheterodyne receivers are provided with at least one r-f amplifier stage.

Also, if the mixer stage were connected directly to the antenna, unwanted signals, called IMAGES, might be received, because the mixer stage produces the intermediate frequency by heterodyning two signals whose frequency difference equals the intermediate frequency. (The heterodyne principle is treated later in this chapter.)

The image frequency always differs from the desired station frequency by twice the intermediate frequency—

$$\text{Image frequency} = \text{station frequency} \pm (2 \times \text{intermediate frequency}).$$

The image frequency is higher than the station frequency if the local oscillator frequency tracks (operates) above the station frequency (fig. 12–8, A). The image frequency is lower than the station frequency if the local oscillator tracks below the station frequency (fig. 12–8, B). The latter arrangement is generally used for the higher frequency bands, and the former, for the lower frequency bands.

For example, if such a receiver having an intermediate frequency of 455 kc is tuned to receive a station frequency of 1500 kc (fig. 12–8, A), and the local oscillator has a frequency of 1955 kc, the output of the i-f amplifier may contain two interfering signals—one from the 1500-kc station and the other from an image station of 2410 kc ($1500+2\times455=2410$ kc). The same receiver tuned near the low end of the band to a 590-kc station has a local oscillator frequency of 1045 kc. The output of the i-f amplifier contains the station signal ($1045-590=455$ kc) and an image signal ($1500-1045=455$ kc). Thus the 1500-kc signal is an image heard simultaneously with the 590-kc station signal.

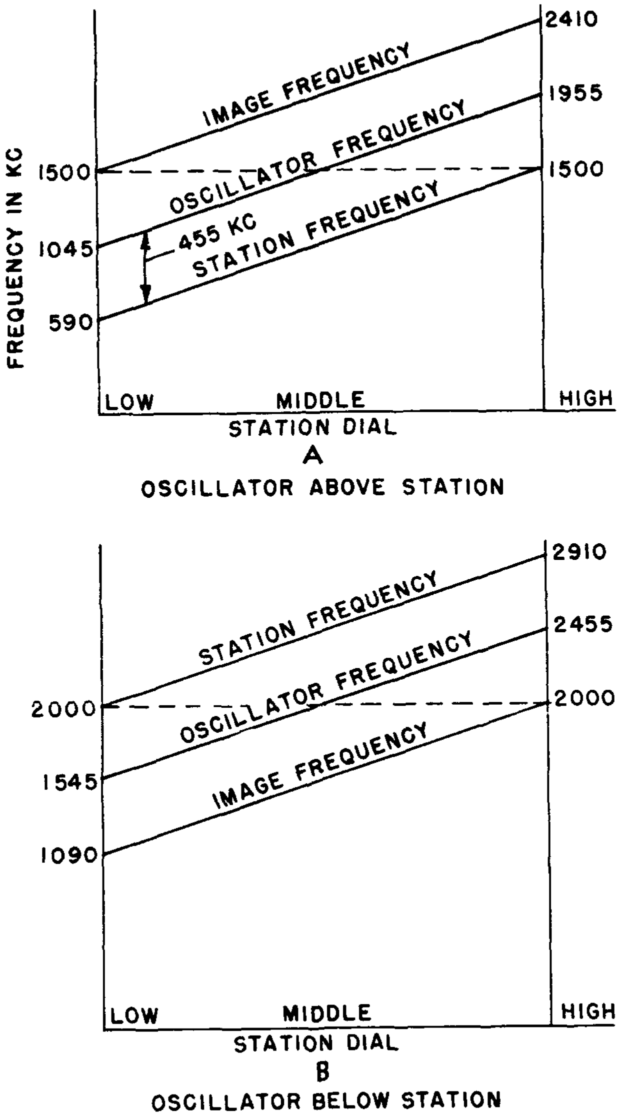


Figure 12-8.—Relation of image frequency to station frequency in a superheterodyne receiver.

It may also be possible for ANY two signals having sufficient strength, and separated by the intermediate frequency, to produce unwanted signals in the reproducer. The selectivity of the preselector tends to reduce the strength of these images and unwanted signals.

The ratio of the amplitude of the desired station signal to that of the image is called the IMAGE REJECTION RATIO and is an important characteristic of a superheterodyne receiver. Better superheterodyne receivers are therefore equipped with one or more preselector stages, a typical example of which is shown in figure 12–9.

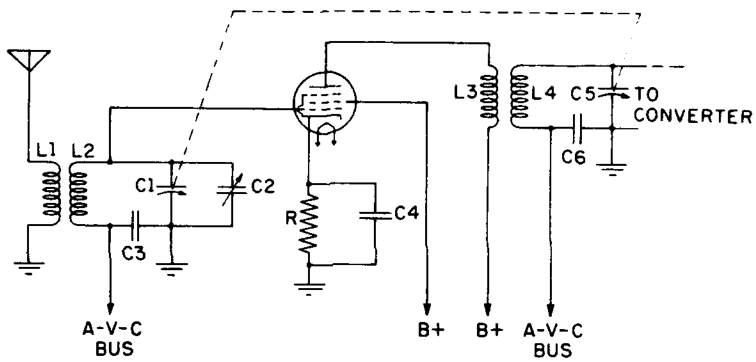


Figure 12–9.—Typical superheterodyne preselector stage.

The preselector stage employs a variable-mu tube and cathode bias. *L*1 is the antenna coil, *L*2 and *C*2 make up the tuned input circuit, and *C*1 is the trimmer used for alignment purposes. The dotted line indicates ganged tuning capacitors. Usually these are the tuning capacitor of the mixer input tank circuit and the local oscillator tuning capacitor. *C*3 provides low impedance coupling between the lower end of *L*2 and the grounded end of *C*2, thus bypassing the decoupling filters in the automatic-volume-control (a-v-c) circuit. (Automatic-volume, or automatic-gain, control is treated later in this chapter.) The r-f transformer in the

output circuit consists of an untuned high-impedance primary, $L3$, and a tuned secondary, $L4$, which resonates with tuning capacitor $C5$ at the station frequency. R-f bypass capacitor $C6$ serves a function similar to that of $C3$.

FIRST DETECTOR.—The first detector, or frequency-converter, section of a superheterodyne receiver is composed of two parts—the oscillator and the mixer. In many receivers, particularly at broadcast frequencies, the same vacuum tube serves both functions, as in the pentagrid converter shown in figure 12–10. The operation of the tube may be simplified

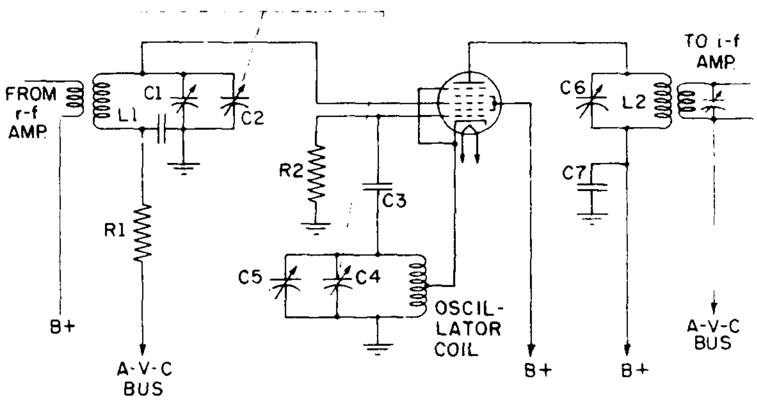


Figure 12–10.—First detector employing a pentagrid converter.

somewhat if both stages (oscillator and mixer) are considered as exerting two different influences on the stream of electrons from cathode to plate. These electrons are influenced by the oscillator stage (grids 1, 2, and 4) and also by the station input signal on grid number 3. Thus, coupling between the input signal and the oscillator takes place within the electron stream itself.

There is a tendency for the local oscillator to synchronize with the station frequency signal applied to grid 3. At high frequencies where the two signals have nearly the same frequency, the pentagrid converter is replaced with a mixer tube and a separate oscillator tube. This type of circuit provides frequency stability for the local oscillator.

The oscillator stage employs a typical Hartley circuit in which $C4$ and the oscillator coil makes up the tuned circuit. $C5$ is the trimmer capacitor which is used for alignment (tracking) purposes. $C3$ and $R2$ provide grid-leak bias for the oscillator section of the tube. Grid 1 is the oscillator grid, and grids 2 and 4 serve as the oscillator plate. Grids 2 and 4 are connected together and also serve as a shield for the signal input grid, 3.

Grid 3 has a variable-mu characteristic, and serves as both an amplifier and a mixer grid. The tuned input is made up of $L1$ and $C2$, with the parallel trimmer $C1$. The dotted lines drawn through $C2$ and $C4$ indicate that both of these capacitors are ganged on the same shaft (in this example with the preselector tuning capacitor). The plate circuit contains the station frequency and the oscillator frequency signals both of which are bypassed to ground through the low reactance of $C6$ and $C7$. The heterodyne action within the pentagrid converter produces additional frequency components in the plate circuit, one of which is the difference frequency between the oscillator and the station frequency. The difference frequency is the intermediate frequency and is amplified by resonance in $C6$ and $L2$. This signal is coupled to the first i-f amplifier grid through the desired band-pass coupling which is wide enough to include the side-band components associated with the amplitude-modulated signal applied to grid 3 of the pentagrid converter.

The conversion gain in a pentagrid converter is,

$$\mu = V_d S_c,$$

where V_d is the a-c plate resistance with the station r-f carrier applied, and S_c is the conversion transconductance (30% to 40% of the g_m of the pentode amplifier). Conversion gain is the change in plate voltage at the intermediate frequency divided by the change in grid voltage at the r-f station frequency for equal changes in plate current at the intermediate frequency. Expressed as a formula,

$$\text{conversion gain} = \frac{\text{i-f output volts}}{\text{r-f input volts}}.$$

The conversion gain of a typical pentagrid converter used in broadcast receivers ranges between 30 and 80.

HETERODYNE PRINCIPLE.—The production of audible beat notes is a phenomenon that is easily demonstrated. For example, if two adjoining piano keys are struck simultaneously, a tone will be produced that rises and decreases in intensity at regular intervals. This action results from from the fact that the rarefactions and condensations produced by the vibrating strings will gradually approach a condition in which they reinforce each other at regular intervals of time with an accompanying increase in the intensity of the sound. Likewise at equal intervals of time, the condensations and rarefactions gradually approach a condition in which they counteract each other, and the intensity is periodically reduced.

This addition and subtraction of the intensities at regular intervals produces BEAT FREQUENCIES. The number of beat frequencies produced per second is equal to the difference between the two frequencies.

The production of beats in a superheterodyne receiver is somewhat analogous to the action of the piano, except that with the receiver the process is electrical and the frequencies are much higher. Figure 12–11 indicates graphically how the beat frequency (intermediate frequency) is produced when signals of two different frequencies are combined in the mixer tube. The resultant envelope varies in amplitude at the difference frequency, as indicated by the dotted lines.

In this example, one voltage, e_s, has a frequency of 8 cycles per second and the other voltage, e_o, has a frequency of 10 cycles per second. Initially, the amplitudes of the two voltages add at instant A, but at instant B the relative phase of e_o has advanced enough to oppose e_s, and the amplitude of the resultant envelope is reduced to a value dependent upon e_s. At instant C the relative phase of e_o has advanced enough to permit the amplitudes to add again. Thus, 1 cycle of amplitude variation of the envelope takes place in the time interval that e_o needs to gain 1 cycle over e_s. From figure 12–11 it may be seen that e_o gains 2 cycles in the interval

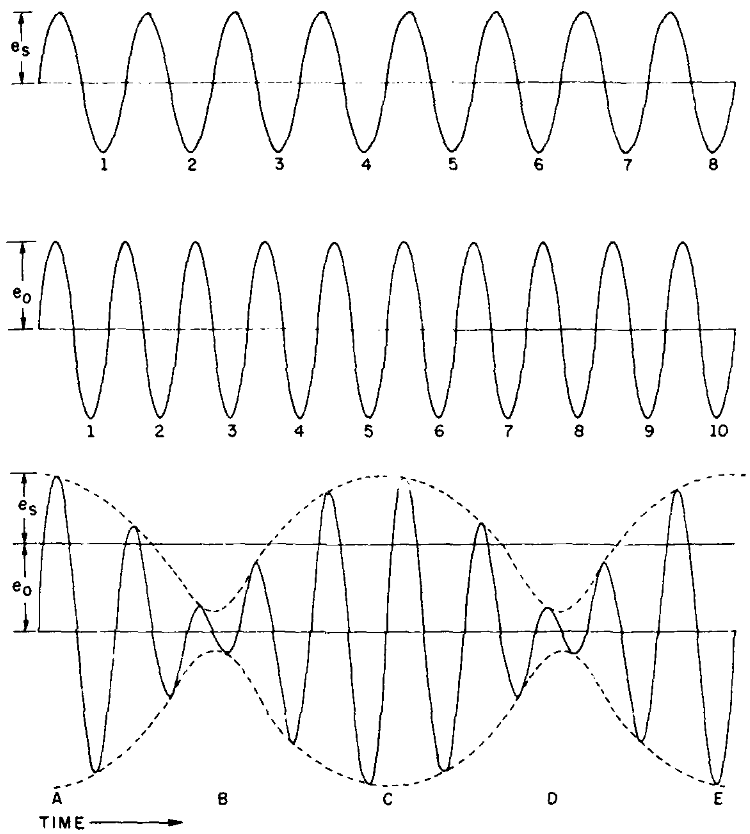


Figure 12–11.—Simplified graphical analysis of the formation of beats.

A to *E*. Therefore, the beat or difference frequency is 2 cycles per second.

I-F AMPLIFIER.—The i-f amplifier is a high-gain circuit commonly employing pentode tubes. This amplifier is permanently tuned to the frequency difference between the local oscillator and the incoming r-f signal. Pentode tubes are generally employed, with one, two, or three stages, depending on the amount of gain needed. As previously stated, all incoming signals are converted to the same fre-

quency by the frequency converter, and the i-f amplifier operates at only one frequency. The tuned circuits, therefore, are permanently adjusted for maximum gain consistent with the desired band pass and frequency response. These stages operate as class-A voltage amplifiers and practically all of the selectivity of the superheterodyne receiver is developed by them.

Figure 12–12 shows the first i-f amplifier stage. The

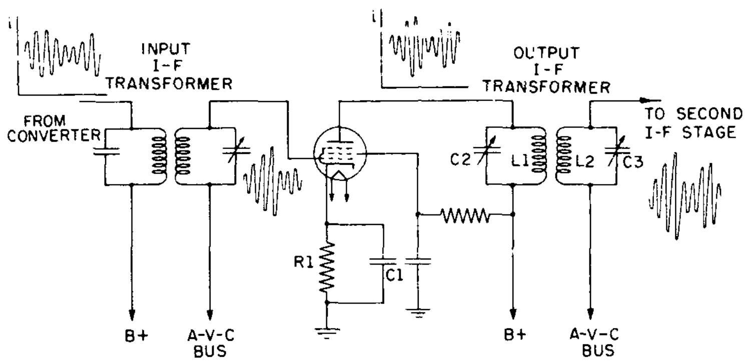


Figure 12–12.—First i-f amplifier stage.

minimum bias is established by means of $R1C1$, and automatic volume control is applied to the grid through the secondary of the preceding coupling transformer.

The output i-f transformer, which couples the plate circuit of this stage to the grid circuit of the second i-f stage, is tuned by means of capacitors $C2$ and $C3$. Mica or air-trimmer capacitors may be used. In some instances the capacitors are fixed, and the tuning is accomplished by means of a movable powdered-iron core. This method is called PERMEABILITY tuning. In special cases the secondary only is tuned. The coils and capacitors are mounted in small metal cans which serve as shields, and provision is made for adjusting the tuning without removing the shield.

The input i-f transformer has a lower coefficient of coupling than the output transformer in some receivers in order to

suppress noise from the pentagrid converter. The output i-f transformer is slightly overcoupled with double humps appearing at the upper and lower side-band frequencies. The over-all response of the stage is essentially flat, and in typical broadcast receivers has a voltage gain of about 200 with a band pass of 7 to 10 kc and an i-f of about 456 kc.

The chief characteristic of the double-tuned band-pass coupling is that at frequencies slightly above and slightly below the intermediate frequency the impedance coupled into the primary by the presence of the secondary is reactive. This cancels some of the reactance existing in the primary, and the primary current increases. Thus the output voltage of the secondary does not fall off and the response is uniform within the pass band. The double-tuned i-f amplifier is discussed in detail in chapter 5.

Crystal filter.—A quartz crystal, used as a selective filter in the i-f section of a communications receiver, is one of the most effective methods of achieving maximum selectivity. It is especially useful when the channel is crowded and considerable noise (both external and internal) is present. The crystal acts as a high-Q tuned circuit, which is many times more selective than tuned circuits consisting of inductors and capacitors. The crystal dimensions are so chosen that the crystal will be in resonance at the desired intermediate frequency.

One of the simplest of a number of possible circuit arrangements is shown in figure 12–13. The crystal is in one arm of a bridge circuit. The secondary of the input transformer is balanced to ground through the center-tap connection. The phasing capacitor, $C4$, is in another arm of the bridge circuit. The crystal acts as a high-Q series-resonant circuit and allows signals within the immediate vicinity of resonance to pass through the crystal to the output coil, $L3$. The desired signal appears between the center tap of $L3$ and ground.

The capacity between the crystal holder plates may bypass unwanted signals around the crystal. Therefore, some method must be provided to balance out this capacitance.

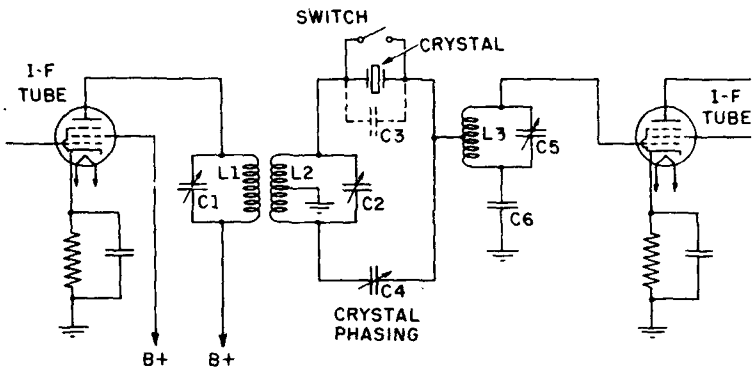


Figure 12-13.—Crystal filter used in the i-f section of a superheterodyne receiver.

In this circuit, balancing is accomplished by taking a voltage 180° out of phase with the instantaneous voltage across the crystal and applying it via $C4$ in such a way as to neutralize the undesired signal voltage. The balanced input circuit in this case is obtained by the use of a center-tapped inductor. The tap on $L3$ permits the proper impedance match.

SECOND DETECTOR.—Most superheterodyne receivers employ a diode as the second detector. This type of detector is practical because of the high gain as well as the high selectivity of the i-f stages. The diode detector has good linearity and can handle large signals without overloading. For reasons of space and economy, the diode detector and first audio amplifier are often included in the same envelope in modern superheterodyne receivers.

A simple diode detector is shown in figure 12-14. The rectified voltage appears across $R1$, which also serves as the volume-control potentiometer. Capacitor $C2$ bypasses the r-f component to ground, and $C3$ couples the output of the detector to the first audio amplifier stage. The tuned circuit $L2C1$ is the secondary of the last i-f transformer.

The time constant of R_1C_2 is long compared to the time for one i-f cycle but short compared to the time for one a-f

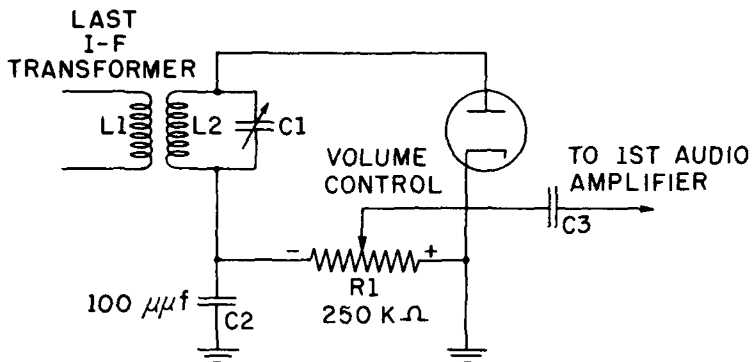


Figure 12-14.—Diode detector.

cycle. If the intermediate frequency is 456 kc the time for one i-f cycle in microseconds is

$$\frac{1}{0.456}=2.19\ \mu s.$$

If R_1 is 250 k-ohms and C_2 is 100 $\mu\mu$f the time constant in microseconds is

$$0.25\times 100=25\ \mu s.$$

The demodulation capacitor, C_2, discharges through R_1 in one-half the time for one a-f cycle $\left(\frac{1}{2f}\right)$. The time required to completely discharge C_2 is $5R_1C_2$ seconds. Thus,

$$\frac{1}{2f}=5R_1C_2$$

$$f=\frac{1}{10R_1C_2}$$

$$=\frac{1}{10\times 0.250\times 10^6\times 100\times 10^{-12}}$$

$$=\frac{10^6}{10_3\times 0.250}=4{,}000\text{ cps.}$$

Thus, the highest audio frequency which C_2 is capable of following without distortion is, in this example, 4,000 cps. In order to increase the response of the diode detector the time constant of R_1C_2 is reduced, for example, by decreasing R_1 to 100 k-ohms. The highest audio frequency now becomes

$$f=\frac{1}{10R_1C_2}$$

$$=\frac{1}{10\times 0.100\times 10^6\times 100\times 10^{-12}}$$

$$=\frac{10^6}{10^2}=10{,}000 \text{ cps.}$$

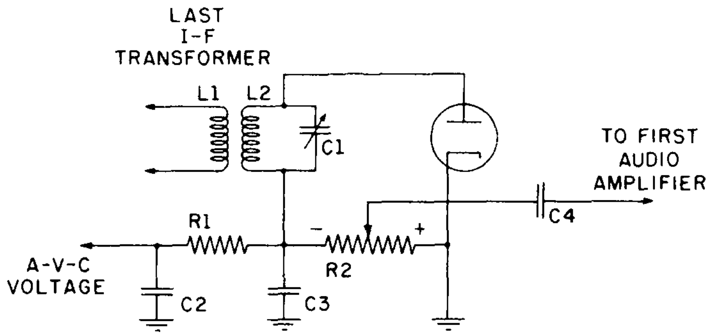
Demodulation capacitor C_2 cannot discharge rapidly enough to follow modulation frequencies higher than 10,000 cps (in this case), and clipping results with all higher audio frequencies. The diode detector is discussed in detail in chapter 8.

AUTOMATIC GAIN CONTROL.—Under ideal conditions, once the manual volume or gain control has been set, the output signal should remain at the same level even if the input signals vary in intensity. The development of variable-mu tubes makes it possible to devise a practical a-v-c or a-g-c circuit, since the amplification of the tube may be controlled by varying the grid-bias voltage. All that is needed is a source of bias voltage that varies with the signal strength. If this voltage is applied as bias to the grids of the variable-mu r-f amplifier stages, the grids will become more negative as the signal becomes stronger. The amplification will thus be reduced, and the output of the receiver will tend to remain at a constant level. Unless the selectivity of the i-f stages is good, strong adjacent-channel signals will reduce receiver gain when a weak signal is tuned in. When no interference is present, a-v-c holds the audio output constant as the input signal amplitude varies over a wide range.

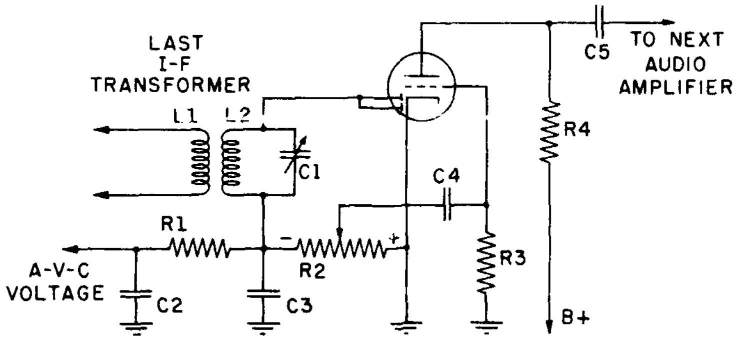
The LOAD RESISTOR of a diode detector is an excellent source of this voltage, since the rectified signal voltage will

increase and decrease with the signal strength. A filter is used to remove the a-f component of the signal and at the same time to prevent the a-v-c circuit from shorting the audio output. Only the slower variations due to fading or change of position of the receiving antenna, and so forth, will then affect the gain of the r-f amplifier stages.

Figure 12–15 shows how the a-v-c voltage is obtained.



A
DETECTOR, A-V-C, AND MANUAL VOLUME CONTROL CIRCUITS



B
DETECTOR, A-V-C, FIRST AUDIO AMPLIFIER, AND MANUAL VOLUME CONTROL CIRCUITS.

Figure 12–15.—Manual and a-v-c circuits.

The a-v-c voltage is tapped off at the negative end of the diode load resistor, $R2$ (fig. 12–15, A), which is also the manual volume control. The a-f component is removed by the filter circuit that is composed of $C2$ and $R1$. One or more of the r-f amplifiers may be controlled in this manner. A customary value for $R1$ is 2 megohms and for $C2$ is 0.05 μf.

Figure 12–15, B, shows an a-v-c circuit used with a duodiode triode in a conventional diode detector circuit. The two plates of the diode are connected together to form a half-wave rectifier in the r-f portion of the circuit. The output of the diode detector is fed to the grid of the triode section which acts as a class-A voltage amplifier.

The variable-mu tube is designed to operate with a minimun bias of about −3 volts. The minimum bias is usually provided by a cathode resistor, and the a-v-c bias is applied in series with it. A disadvantage of ordinary automatic volume control is that even the weakest signals produce some a-v-c bias, which reduces the amplification slightly.

Delayed automatic gain control.—The disadvantage of automatic gain control, that of attenuating even the very weak signals, is overcome by the use of delayed automatic

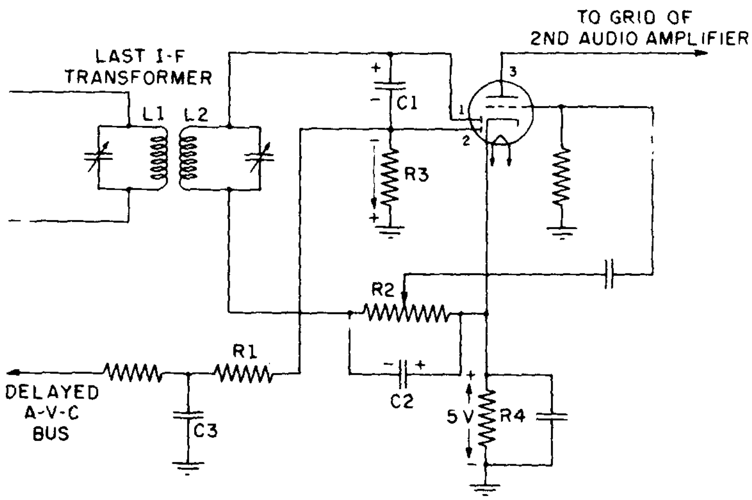


Figure 12–16.—Delayed a-v-c circuits.

gain control, as shown in figure 12–16. In this circuit the a-v-c diode, plate 2, is separated from the detector diode, plate 1, and both are housed in the same envelope with a triode amplifier.

In this example a bias of 5 volts on the delayed a-v-c diode, plate 2, prevents it from conducting until the signal exceeds 5 volts. The signal across the secondary of the i-f transformer is coupled to diode, plate 2, by capacitor $C1$. Until the signal exceeds 5 volts no charge is acquired by the a-v-c capacitor, $C3$; no additional bias is applied to the grids of the i-f amplifier, preselector, or converter tubes; and their gain is maximum on weak signals. The 5-volt bias applied to the delayed a-v-c diode, plate 2, is developed across cathode resistor $R4$ by the current flowing through the triode section of the tube. The triode section serves as a class-A voltage amplifier driven by the audio voltage developed across diode load resistor $R2$.

When the signal across the secondary of the i-f transformer exceeds the 5-volt bias value across $R4$, the a-v-c diode (plate 2) conducts on alternate half cycles and $C3$ acquires a charge. The voltage developed across $C3$ constitutes the delayed a-v-c voltage. It is supplied to the grids of the various stages ahead of the second detector in series with the cathode bias developed by the individual tubes.

Noise limiter.—Sudden bursts of noise in a receiver may be attenuated by the use of a series noise limiter, such as the one shown in figure 12–17. The diode detector circuit includes $V1$, $R1$, $R2$, and $C2$. The cathode of $V2$ is connected through $R3$ to the a-v-c line, which is negative with respect to ground. The plate of $V2$ is connected to the common connection, B, between the diode load resistors, $R1$ and $R2$.

When a normal signal is detected, the plate of $V2$ is negative with respect to ground by an amount equal to the voltage drop across $R1$. Normally the plate is less negative than the cathode, and $V2$ conducts, thus providing a continuous circuit through $V2$, for the audio voltage tapped off at point D.

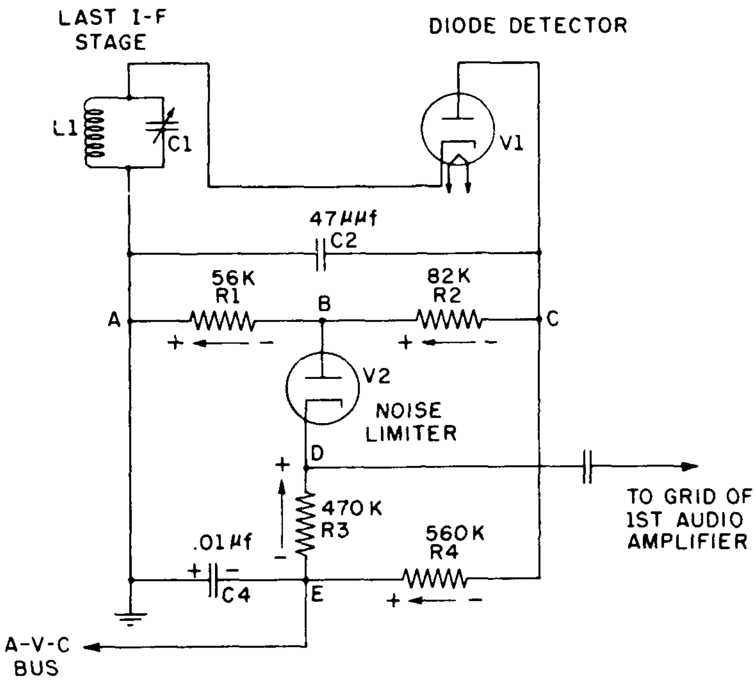


Figure 12-17.—Simplified circuit of a diode detector and series noise limiter.

If a sudden burst of noise comes through the receiver the voltage across $R1$ suddenly increases. A large negative potential with respect to ground is thus applied to the plate of $V2$. The cathode cannot follow this sudden change because of the long time constant of the a-v-c circuit. The plate is now negative with respect to the cathode, and $V2$ ceases to conduct. Thus the output circuit to the audio amplifiers is opened and the receiver becomes momentarily quiet. The point at which $V2$ begins limiting depends on the average strength of the received signal in relation to the amplitude of the noise. On weak signals the a-v-c voltage and cathode bias are both small; thus a low-amplitude noise pulse swings the plate voltage negative with respect to the cathode bias, and $V2$ limits the audio output. Conversely, a strong signal

is limited by a large amplitude noise pulse but is not limited by the same amplitude of noise pulse that cut off the weak signal.

Beat-frequency oscillator.—The beat-frequency oscillator (BFO) is necessary when c-w signals are to be received because these signals are not modulated with an audio component. In superheterodyne receivers the incoming c-w signal is converted to the intermediate frequency at the first detector as a single frequency signal with no side-band components. The i-f signal is heterodyned (with a separate tunable oscillator known as the beat-frequency oscillator) at the second detector to produce an a-f output. In the circuit shown in figure 12–18, the Hartley oscillator (BFO)

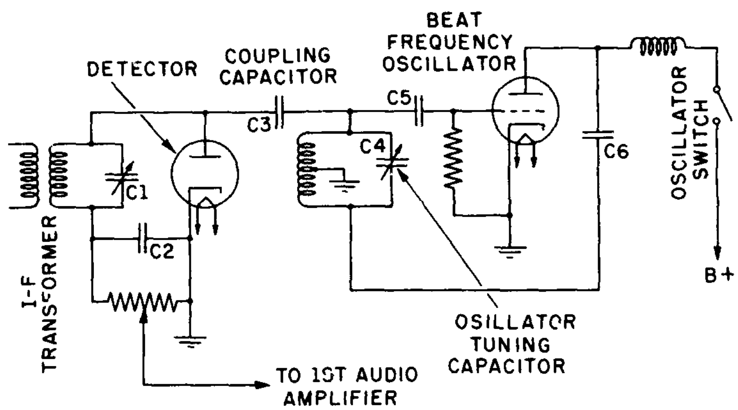


Figure 12–18.—Beat-frequency oscillator.

is coupled to the plate of the second detector by capacitor *C*3.

If the intermediate frequency is 455 kc and the BFO is tuned to 456 kc or 454 kc, the difference frequency of 1 kc is heard in the output. Generally the switch and capacitor tuning control are located on the front panel of the receiver.

The BFO should be shielded to prevent its own output from being radiated and combined with desired signals ahead of the second detector. If a-v-c voltage is to be used it should be obtained from a separate diode isolated from the second detector. One way is to couple the output of an i-f amplifier

stage ahead of the second detector to the a-v-c diode. Otherwise, the output of the BFO would be rectified by the second detector and would develop an a-v-c voltage even on no signal.

SILENCE.—A silencer is sometimes employed in the a-f section of a receiver to disable the receiver when no signals are being received. One type of silencer circuit is shown in figure 12-19.

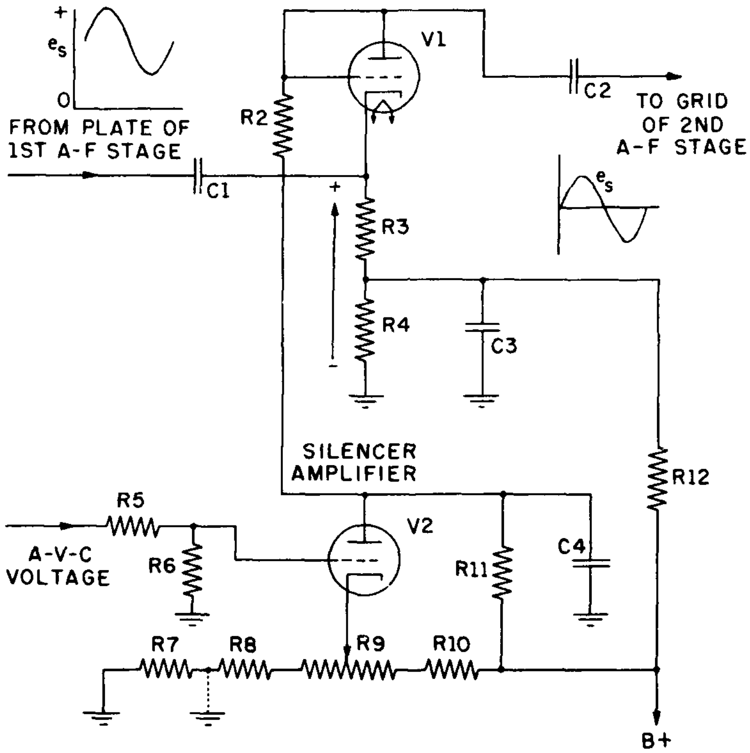


Figure 12-19.—Silencer circuit.

The silencer, $V1$, a diode-connected triode, connects the output of the first a-f stage to the input of the second audio amplifier. Silencer amplifier $V2$ serves as the control tube

for the silencer. The plate voltage of $V1$ is supplied via $R2$ from the plate of $V2$ (which is in turn supplied from the B supply via $R11$) and is positive with respect to ground. The cathode voltage of $V1$ is also positive with respect to ground, since it is connected to the B supply through a voltage divider made up of $R12$ and $R4$. With no input signal, $R9$ is adjusted until $V2$ draws enough plate current to reduce its plate voltage and that of $V1$ to a value below the voltage on the cathode of $V1$. Thus the silencer plate voltage is negative with respect to the cathode. Conduction ceases, and the silencer cuts off. The output is reduced to zero, and the receiver is mute.

The grid of $V2$ is connected to the a-v-c line. When a signal enters the receiver, the negative a-v-c voltage is applied to the grid of $V2$, thereby reducing the plate current and increasing the plate voltage of both $V2$ and $V1$. When the plate of $V1$ becomes positive with respect to its cathode, the tube conducts and the signal is passed to the second a-f amplifier.

Circuit of a Superheterodyne Receiver

The complete circuit of a superheterodyne receiver is shown in figure 12–20. In this circuit one r-f amplifier (preselector) stage is used. Tube $V2$, a pentagrid converter, serves both as the mixer tube and oscillator tube. Three tuning capacitors (one each in the preselector, mixer, and oscillator stages) are ganged on a common shaft to assure proper tracking. Trimmers are connected in parallel with each tuning capacitor to permit alinement. The oscillator tuning capacitor is smaller than the tuning capacitor in the preselector or the converter stages. The oscillator operates above the station frequency and tracks closely at three points on the dial— (1) low end, (2) middle, and (3) high end. The oscillator tuning capacitor split-rotor plates allow closer adjustment for tracking at the low end and at the middle of the band. Shunt trimmer capacity adjustments on the

oscillator tuning capacitor provide close tracking of the oscillator at the high end of the band.

Tube *V3* is the i-f amplifier with input and output i-f transformers tuned to the receiver intermediate frequency.

Tube *V4* serves as the second detector and first audio amplifier. Conventional automatic volume control is tapped off at the end of the volume control potentiometer farthest from ground. Plate and screen potentials are obtained from the B supply through the corresponding voltage dropping resistors. The power supply is a conventional full-wave rectifier.

F-M RECEIVERS

The t-r-f and superheterodyne receivers that have been described in the preceding paragraphs of this chapter are designed to receive r-f signals that vary in amplitude according to the audio modulation at the transmitter. The amplitude of the r-f signal is increased by one or more r-f amplifier stages, and the modulation component is removed by the detector. Each of the tuned circuits preceding the detector is designed to pass only a relatively narrow band of frequencies containing the necessary upper and lower sideband frequencies associated with the amplitude-modulated carrier.

F–m receivers are supplied r-f signals that vary in frequency according to the information being transmitted. The amount of the variation or deviation from the CENTER, or RESTING, FREQUENCY at a given instant depends on the amplitude of the impressed audio signal. The frequency with which the variations from the center frequency occur depends on the frequency of the impressed audio signal. The function of the f-m receiver is basically the same as that of the a-m superheterodyne receiver—that is, the amplitude of the incoming r-f signals is increased in the r-f stages; then the frequency is reduced in the mixer stage to the intermediate frequency and amplified in the i-f amplifier section. Finally, the amplitude is clipped in the limiter stage and the modula-

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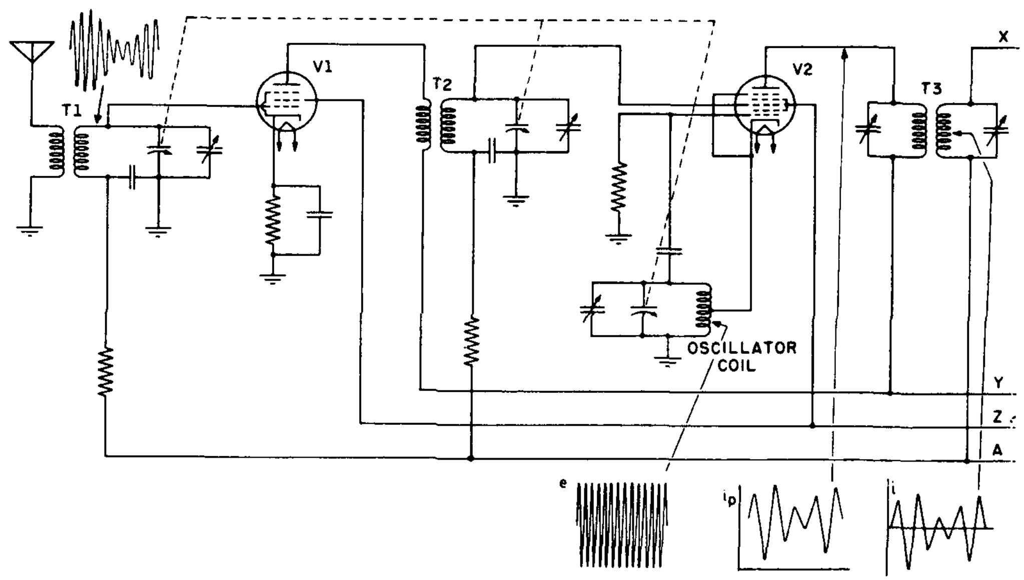

Figure 12-20.—Circuit diagram of a superheterodyne receiver.

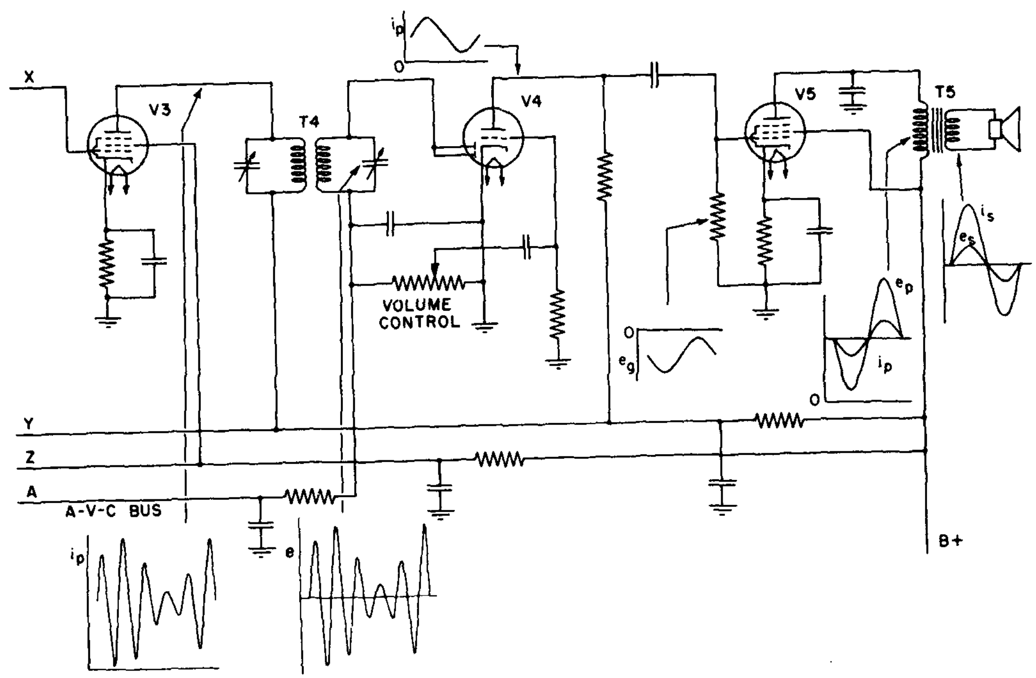

Figure 12-20.—Circuit diagram of a superheterodyne receiver—Continued

tion component is removed by the second detector, or DISCRIMINATOR as it is called in the f-m receiver.

There are a few major differences between the f-m and the a-m receiver. The greatest difference is in the method of detection. Also the tuned circuits of the f-m receiver have wider pass bands, and the last i-f stage is especially adapted for limiting the amplitude of the incoming signal. However, in both systems the audio amplifiers and reproducers are similar.

A comparison between a superheterodyne receiver designed for a-m reception and one designed for f-m reception is shown in figure 12–21.

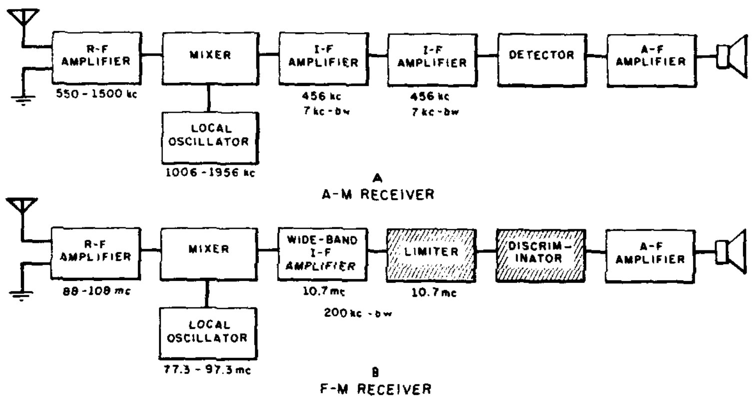

Figure 12–21.—Receiver block diagrams.

Components

R-F SECTION.—The function of the f-m antenna is to provide maximum signal voltage to the receiver input. Unlike most broadcast a-m receiver antennas, f-m receiver antennas act as resonant lines having standing waves on them. Therefore f-m antennas are cut to the required length in order to receive a signal of sufficient amplitude to drive the first r-f amplifier. If a single frequency is to be received, the antenna may be designed for maximum response

at that frequency. If, however, a band of frequencies is to be received, the antenna length will represent a compromise. Usually the length is so chosen that it will be in resonance at the geometric center of the band. The GEOMETRIC CENTER or MEAN is equal to $\sqrt{\lambda_1 \lambda_2}$, where λ_1 and λ_2 are the wavelengths at the two ends of the band.

There are many types of f-m antennas, but probably the simplest is the half-wave dipole. The length of the half-wave dipole, in feet, is

$$\frac{468}{\sqrt{f_1 f_2}},$$

where f_1 and f_2 are the frequencies in megacycles at the two ends of the band. Because the resistance at the center of the half-wave dipole is about 72 ohms, the transmission line connecting the antenna with the receiver should have a characteristic impedance of 72 ohms in order to operate as a nonresonant transmission line with no standing waves. The transmission line feeds the signal to the receiver via a matching transformer at the input to the preselector stage.

The r-f amplifier, or preselector, performs essentially the same function in the f-m receiver as it does in the a-m receiver—that is, it increases the sensitivity of the receiver. Such an increase in sensitivity is often a practical necessity in fringe areas. However, the gain of the i-f stages is relatively much greater, perhaps 10 times that of the preselector, since the chief advantage of the superheterodyne lies in the uniformity of response and gain of the i-f stages within the receiver band. The principal functions of the r-f stage are to discriminate against undesired signals (images) and to increase the amplitude of weak signals so that the signal-to-noise ratio will be improved.

If the receiver is designed to receive both amplitude modulation and frequency modulation, a suitable band-switching arrangement is necessary. Many combination receivers are designed to receive more than one a-m band. Under such circumstances additional tuned circuits are needed. Thus, if two a-m bands and one f-m band are used

and one r-f stage is used ahead of the mixer, three tuned circuits are needed for each band to be covered. This circuit arrangement includes one for the r-f stage, one for the mixer stage, and one for the oscillator stage for each of the three bands, or a total of nine tuned circuits. The f-m tuned circuits have wider band-pass characteristics than do the a-m tuned circuits, as shown in figure 12-21.

FREQUENCY CONVERTER.—The frequency converter employed in the f-m receiver functions in much the same manner as the one employed in the a-m superheterodyne receiver. However, additional problems are involved.

For example, at the frequencies employed in the commercial f-m band the stability of the local oscillator becomes a major problem. As mentioned before, there is a tendency for the local oscillator to become synchronized with the incoming signal and thus to lose the intermediate frequency output entirely. The tendency is more pronounced at f-m frequencies because the station and oscillator are closer together. Therefore, for maximum frequency stability, separate oscillator tubes are used. This results in increased space requirements and expense. Especially designed pentagrid converters that have reasonably good frequency stability, high conversion transconductance, and oscillator transconductance are employed in some less expensive commercial sets.

Even in a normal well-designed f-m receiver such factors as the change in internal capacitance of the oscillator tube (or oscillator section of a tube) and the expansion of coil windings and capacitor plates during warm-up may cause the local oscillator frequency, and consequently the intermediate frequency, to drift an appreciable amount. A relatively small shift in oscillator frequency (always downward with respect to the center frequency in an uncompensated oscillator circuit) may shift the i-f signal beyond the range of the i-f stages with a consequent loss in output signal.

Various methods are used to combat oscillator drift. For example, the second harmonic of the local oscillator frequency is sometimes used for mixing. In this instance the local

oscillator may be operated at a lower fundamental frequency, where the stability is improved. Another method is to use capacitors having a negative temperature coefficient. These are connected in shunt with capacitors having a positive temperature coefficient to counteract the change in capacitance when the temperature of the oscillator stage varies. Proper voltage regulation as well as the choice of oscillator tubes having low internal capacitances, will also increase the stability of the local oscillator.

Frequency stability of the local oscillator, in the standard f-m band, makes it advantageous to operate the local oscillator at a frequency below that of the incoming signal. (See fig. 12-21.)

However, if the local oscillator is operated above the frequency of the incoming signal it is not so likely to interfere with television receivers in the same vicinity that are operating on the lower video channels. Therefore, some commercial f-m receivers have local oscillators operating above the incoming signal.

I-F AMPLIFIER.—The i-f amplifier in an f-m receiver is usually tuned to a center frequency of from 8 to 10 megacycles. It generally employs double-tuned transformers having equal primary and secondary inductances. The band pass is from 150 to 200 kc. The last one or two i-f stages function as a limiter.

The gain of each wide band i-f stage is considerably less than that of the narrow-band a-m. i-f amplifier. Therefore, f-m receiver employs more i-f stages than a corresponding a-m receiver. The gain of an i-f amplifier employing a double-tuned transformer has been given in chapter 5 as

$$\text{V.G.}=\frac{g_m\omega K\sqrt{L_pL_s}}{K^2+\frac{1}{Q_pQ_s}}.$$

In the case of a wide-band i-f amplifier having a double-tuned transformer, critical coupling $\left(K=\frac{1}{Q}\right)$, and primary and

secondary inductances and Q's that are equal, the gain becomes

$$\text{V.G.} = \frac{g_m \omega L Q}{2}.$$

In this formula g_m is the transconductance of the tube, and ωL is the inductive reactance of the circuits at the intermediate frequency.

A low value of intermediate frequency is undesirable because local oscillator drift might force the set to operate outside the i-f range. Also, it would be pointless to have the intermediate frequency lower than the total frequency deviation (bandwidth) of any one f-m station.

In the choice of the optimum i-f value such factors as image response, response to signals at the same frequency as the intermediate frequency, response to beat signals produced by two stations separated in frequency by the i-f value, and response to harmonic frequencies must be considered.

Two stations separated in frequency by the i-f value will, if sufficiently powerful, produce a beat frequency that will pass through the receiver. This type of interference may be eliminated if the intermediate frequency chosen is greater than the entire f-m bandwidth. It may be minimized by adequate discrimination in the preselector stage.

Harmonics of the local oscillator may combine with harmonics produced when a strong incoming signal overloads the input stage to produce the intermediate frequency.

Interf ring signals may develop as a result of the interaction of these harmonic frequencies. For example, consider an f-m receiver having an intermediate frequency of 9.1 mc, and tuned to an 86-mc station. The oscillator frequency is $86+9.1$, or 95.1 mc. It is possible that a strong 90.5-mc signal picked up at the f-m receiver input, would develop at that point its second harmonic of 181.0 mc. The oscillator second harmonic frequency is 95.1×2, or 190.2 mc. The difference frequency is $190.2-181.0$, or 9.20 mc. This difference frequency would appear in the

output of the mixer stage and be accepted by the i-f amplifiers tuned to 9.1 mc. Thus the receiver output would contain the 86-mc station and simultaneously the 90.5-mc interfering signal.

Harmonics produced at the input may be reduced by increasing the selectivity of the tuned circuits and using variable-mu tubes that do not overload easily. The production of harmonics by the local oscillator may be reduced by maintaining a satisfactorily high circuit Q and by reducing its loading.

The uniform gain within the band pass of an a-m i-f amplifier, and the selectivity of i-f amplifiers, are treated in chapter 5. In commercial f-m i-f amplifiers the band pass is considerably greater than it is in a-m i-f amplifiers because of the greater frequency swing used in frequency modulation. An ideal frequency response curve is difficult to obtain economically. Therefore, a practical compromise that gives the necessary uniform gain and discrimination against adjacent channel frequencies is chosen.

The i-f stage may be designed for f-m only or for both a-m and f-m. An i-f transformer designed for both a-m and f-m is shown in figure 12–22. In order to have the desired high L/C ratio for increased gain and increased bandwidth, permeability tuning is employed. Circuits $C1L1$ and $L2C2$ are tuned to the higher f-m intermediate frequency, about 10 mc, and have greater band pass, about 200 kc. Circuits $C3L3$ and $L4C4$ are tuned to the lower a-m intermediate frequency, perhaps 455 kc, and the band pass is lower, about 7 kc.

When the receiver is adjusted for f-m reception, only the f-m section of the i-f transformer is effective in coupling signal voltage to the next tube. Capacitor $C3$, having a low reactance to the higher f-m signals, shunts the a-m section of the transformer. Likewise, when the receiver is adjusted for a-m reception, only the a-m section of the i-f transformer is effective in coupling signal voltage to the next tube. In this case $L1$ becomes an effective short circuit for the lower frequency a-m signals.

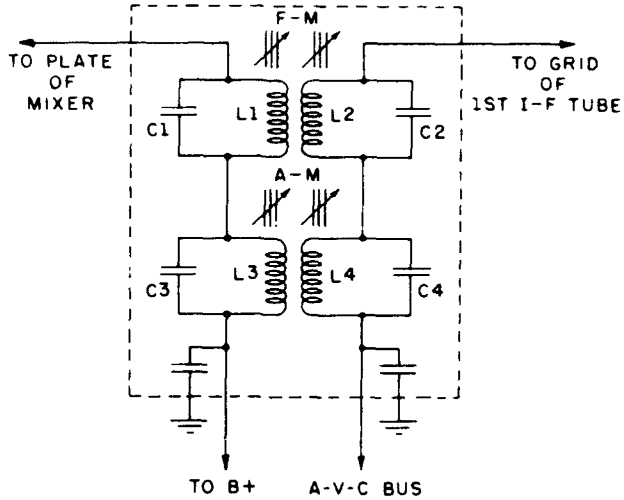

Figure 12-22.—I-f transformers for a-m and f-m.

Usually the last i-f stage is modified to operate as a limiter.

LIMITER.—The limiter in an f-m receiver removes amplitude modulation and passes on to the discriminator an f-m signal of constant amplitude.

As the f-m signal leaves the transmitting antenna it is varying in frequency according to an audio-modulating signal, but it has essentially a constant amplitude. As the signal travels between the transmitting and receiving antenna, however, natural and man-made noises, or static disturbances, are combined with it to produce variations in the amplitude of the modulated signal. Other variations are caused by fading of the signal. Fading might be caused, for example, by movement of the ship carrying the transmitter or the receiver. Still other amplitude variations are introduced within the receiver itself because of a lack of uniform response of the tuned circuits.

All of these undesirable variations in the amplitude of the f-m signal are amplified as the signal passes through the

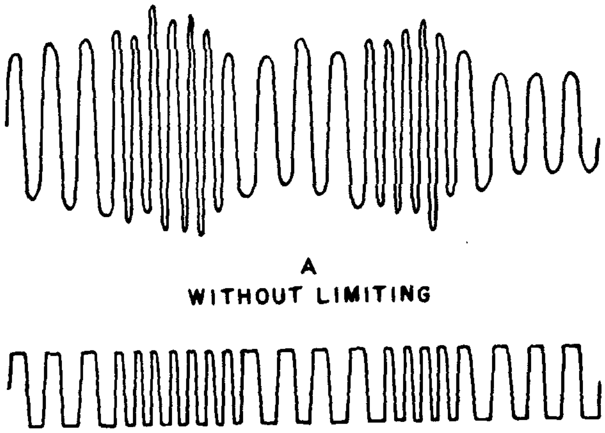

Figure 12–23.—F-m signals.

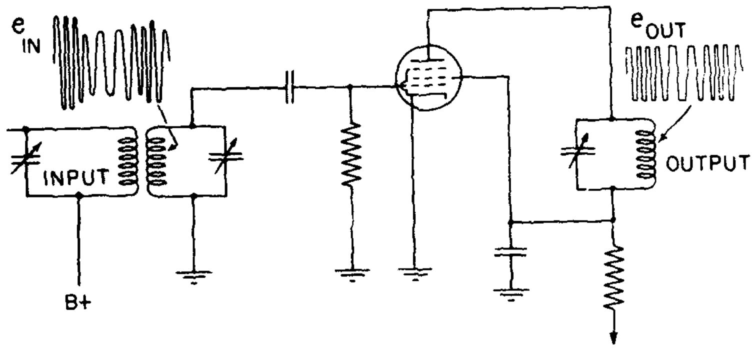

Figure 12–24.—Grid-leak bias limiter.

successive stages of the receiver up to the input of the limiter. This condition in which both frequency modulation (desired) and amplitude modulation (undesired) are present at the same time is shown in figure 12–23, A.

The character of the signal after leaving the limiter should be as indicated in figure 12–23, B, in which all amplitude variations have been removed, leaving a signal that varies only in frequency.

A grid-leak bias limiter is shown in figure 12–24. The tube is a sharp-cutoff pentode operated with grid-leak bias. Because the plate and screen voltages are purposely made low, plate-current saturation as well as plate-current cutoff, is produced readily by input signals having a magnitude of only a few volts.

The manner in which the limiter functions is illustrated by the i_p-e_g curve shown in figure 12–25. Grid-leak bias is used so that with varying signal amplitudes, the bias can adjust itself automatically to a value that allows just the

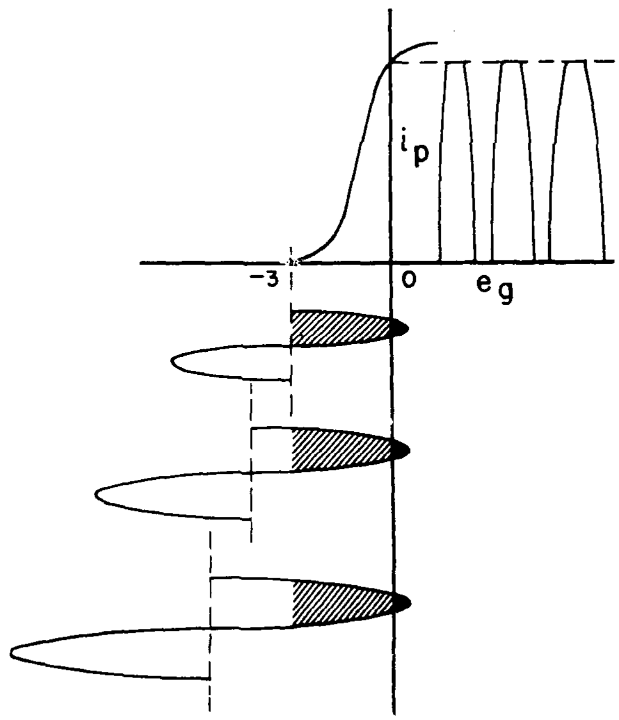

Figure 12–25.—Limiter i_p-e_g curve.

positive peaks of the signal to drive the grid positive and cause grid current to flow.

Suppose that a signal having a peak amplitude greater than the cutoff bias is impressed on the grid of the tube. A bias voltage having a magnitude approximately equal to the peak value of the signal will be developed. Accordingly, grid current will flow for a very small part of the positive half cycle at the peak of signal swing, as shown by the shaded area. Plate current flows for almost the entire positive half cycle. When the signal amplitude increases, a greater bias is developed, but the grid cutoff voltage remains the same and the average plate current changes very little. Thus, the amount of plate-current flow in the limiter stage is approximately constant for all signals having an amplitude great enough to develop a grid-leak bias voltage that is greater than the cutoff voltage. The frequency variations in the f-m signal are maintained in the output because the plate-current pulses are produced at the signal frequency and excite the plate-tuned tank circuit which has a relatively low Q and a wide band pass. Thus, because of the "flywheel" effect, a complete a-c waveform is passed to the secondary of the discriminator transformer for each cycle of input signal.

When the peak amplitude of the grid signal is less than the cutoff voltage, the limiting action fails because the stage is practically a class-A amplifier for such signals, and the average plate current varies as the grid-leak bias changes with varying signal amplitudes. For this reason, the stages preceding the limiter must have sufficient gain to provide satisfactory limiting action on the weakest signal to be received.

DISCRIMINATOR.—Another major difference between the a-m receiver and the f-m receiver is in the method used to detect the signal. The DETECTOR in an a-m receiver interprets the AMPLITUDE VARIATIONS of the amplitude-modulated r-f energy in terms of the audio signal. In the f-m receiver, the discriminator interprets the FREQUENCY VARIATIONS of the frequency-modulated r-f energy in terms of the audio signal.

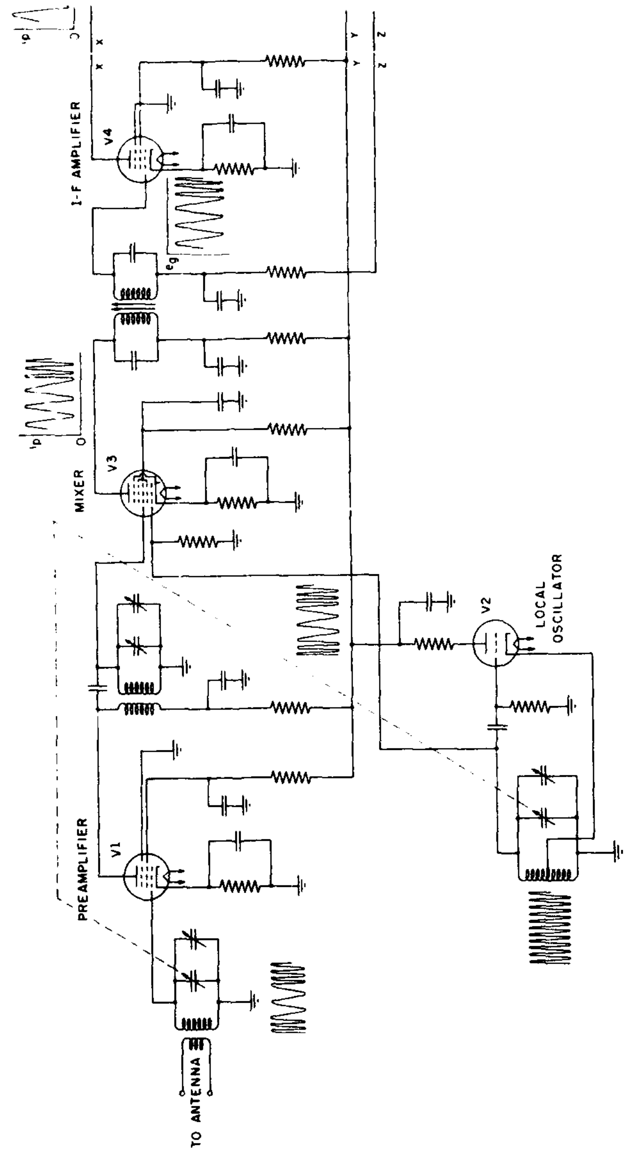

Figure 12-26.—F-m tuner.

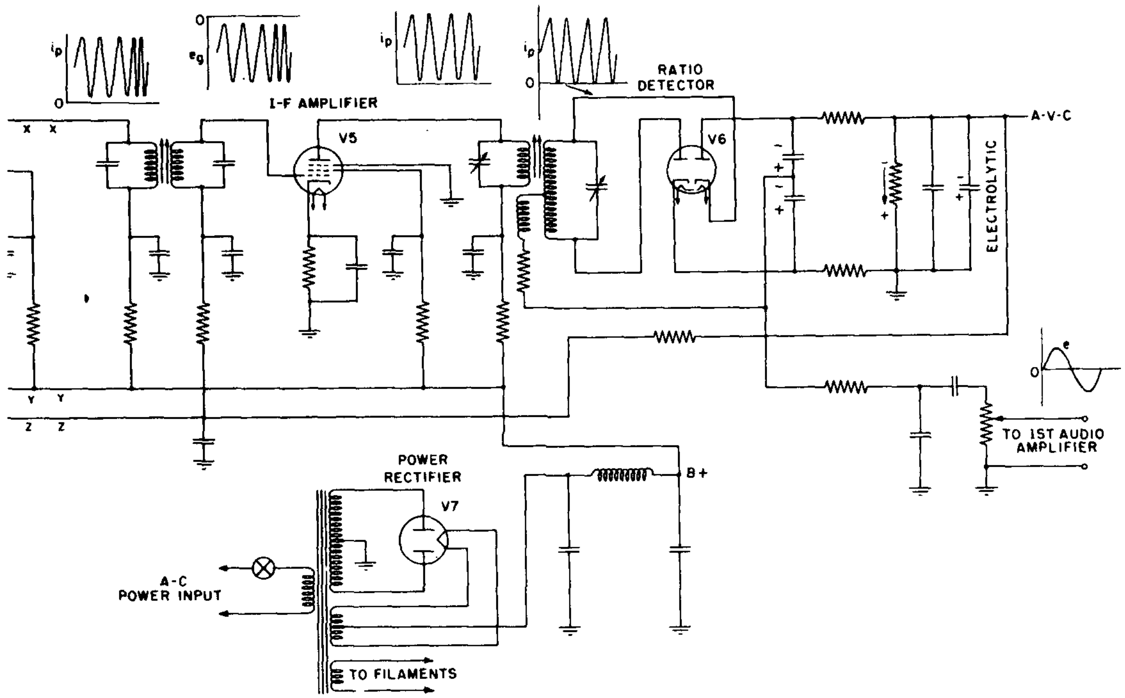

Figure 12-26.—F-m tuner—Continued.

Several types of f-m detectors have been developed and are in use, but perhaps two of the most common types are the discriminator and the ratio detector. The operation of one type of each of these detectors is treated in chapter 8.

The discriminator requires a limiter, which in turn requires considerable amplification ahead of its input.

An f-m detector that would be insensitive to amplitude variations would eliminate the need for a limiter, and in addition one or more i-f amplifier stages might be eliminated. Such an improved discriminator circuit that meets these requirements to a larger degree than the discriminator, is the RATIO DETECTOR.

The principal difference between the ratio detector and the Foster-Seeley discriminator lies in the method of connecting the two rectifier diodes. In the discriminator the diode plates are connected to the ends of the transformer secondary. In the ratio detector the diodes are connected in series across the secondary. The construction of the transformer is also different. The ratio detector has an additional winding called a LINK OR TERTIARY WINDING. Its function is the same as that of the coupling capacitor used in the discriminator—that is, to introduce a voltage differing in phase from the signal voltage developed across the upper and lower halves of the transformer secondary.

Circuit of an FM Tuner

A schematic diagram of an f-m tuner is shown in figure 12–26. Tubes *V*1 and *V*3 are remote-cutoff tubes using cathode bias without automatic volume control. Automatic volume control is not so important in f-m as it is in a-m, since in f-m, particularly if the second detector is a discriminator, the i-f stages are operated at maximum gain. The ratio detector shown in the figure provides a convenient source of a-v-c voltage, which is supplied to the grids of *V*4 and *V*5. The tuning range of the input tank, and also that of the tuned circuit in the mixer input, is 88 to 108 megacycles.

If the intermediate-frequency stages are tuned to 10 mc,

and the local oscillator, $V2$, is operated above the station frequency, then the local oscillator is tunable from 88+10, or 98 mc, to 108+10, or 118 mc. Oscillator tube $V2$ is especially designed for high-frequency operation. Tube $V3$ is a pentagrid mixer used for mixing the incoming signal with the locally generated signal.

The i-f amplifiers are remote-cutoff pentodes. As mentioned previously, the i-f transformers must have the desired wide band-pass characteristic (200 kc).

Tube $V6$, a twin diode, is operated as a ratio detector. The audio output from this detector is fed to a conventional audio amplifier, not shown in the circuit diagram.

The B supply is obtained from a full-wave rectifier, $V7$, as shown in figure 12–26.

QUIZ

1. Distinguish between selectivity and sensitivity of a receiver.
2. What are two reasons tetrodes or pentodes are generally used in r-f amplifiers?
3. When a common B supply is used in a multistage amplifier, why does the greatest amount of feedback occur between the final and first amplifier stages?
4. What are *R-C* circuits called that are designed specifically to counteract feedback in both r-f and a-f amplifiers?
5. What is the function of so-called electrical or mechanical band-spread used in receiver tuning?
6. What are two functions of an a-m detector?
7. What portion of the i_p-e_g curve, on which the plate detector is operated, accounts for the introduction of some distortion?
8. In most cases, what determines the amount of amplification required in the a-f section of a receiver?
9. What is the relative magnitude of the secondary impedance compared with that of the voice coil?
10. In a permanent-magnet dynamic type of loudspeaker, to what electrical quantity is the force on the voice coil proportional?
11. What is the principal disadvantage of a t-r-f receiver?
12. What are two important functions of the r-f amplifier in a superheterodyne receiver besides that of amplifying the signal?

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13. What is meant by an image frequency?
14. What is the purpose of C3 in figure 12–9?
15. In a pentagrid converter, how does the coupling take place between the input signal and the local-oscillator signal?
16. At high frequencies, why is the pentagrid converter often replaced by a mixer tube and a separate oscillator tube?
17. What are two functions performed by grid number 3 in the pentagrid converter of figure 12–10?
18. Conversion gain expresses a ratio between what two voltages?
19. What class of amplifier is used in the i-f stages of a superheterodyne receiver?
20. Why are crystal filters frequently used in the i-f section of communications receivers?
21. Why do most a-m broadcast superheterodyne receivers employ a diode as the second detector?
22. How may the frequency response of the diode detector shown in figure 12–14 be improved?
23. What type of r-f amplifier tubes are necessary if automatic gain control is used?
24. In figure 12–15, how is the automatic-gain-control voltage obtained?
25. (a) What is the approximate minimum bias that is used with variable-mu tubes and (b) how is it obtained?
26. How does delayed a-g-c differ from ordinary a-g-c?
27. In a series noise limiter, when does the limiting action take place—when the limiter diode acts as an open circuit or when it conducts?
28. What is the purpose of the beat-frequency oscillator in a communications receiver?
29. What is the purpose of the silencer used in some communications receivers?
30. What is the principal difference between a-m and f-m receivers?
31. What are the principal functions of the r-f stage in an f-m receiver?
32. How may oscillator drift be reduced in f-m receivers?
33. Why does an f-m receiver employ more i-f stages than a corresponding a-m receiver?
34. What is the function of the limiter in an f-m receiver?

CHAPTER 13

ELECTRONIC TEST EQUIPMENT

CATHODE-RAY OSCILLOSCOPE

Cathode-Ray Tube

The cathode-ray tube is a special type of vacuum tube in which electrons emitted from the cathode are shaped into a narrow beam and accelerated to a high velocity before striking a phosphor-coated viewing screen. The screen fluoresces or glows at the point there the electron beam strikes it and thus provides a visual indication. The cathode-ray tube provides a visual means of examining and measuring current and voltage waveforms. The CATHODE-RAY OSCILLOSCOPE is a test instrument employing the cathode-ray tube. The oscilloscope is one of the most important units of test equipment in maintenance and servicing work. It is used to give a visual presentation of circuit waveforms which, by comparison, show the operating efficiency level of a portion of a circuit, a complete circuit, or an equipment.

The comparison is made against optimum-efficiency waveforms permanently printed, and located either at the equipment test points or on the schematic diagrams. Scope patterns periodically taken at the test points are compared with these printed waveforms. Differences between the optimum waveform and the scope pattern indicate that a circuit (and therefore the equipment) is falling below the

optimum performance level and that corrective action should be applied. By using the oscilloscope in this manner, difficulties may be pin-pointed to a specific circuit or portion of a circuit in a minimum of time.

The tube is also used as the visual indicating device for display information obtained by radar, sonar, radio, direction finders, loran, and television.

The beam of electrons has practically no weight or inertia and follows a straight line unless diverted by an electric or a magnetic field. Cathode-ray tubes are either of two types according to the method of deflecting the electron beam—(1) electrostatic and (2) electromagnetic. The electrostatic type of cathode-ray tube is used in practically all cathode-ray oscilloscopes operating as test instruments. Certain radar and sonar sets use cathode-ray tubes that employ electromagnetic deflection. Focusing or narrowing the beam before deflection is accomplished either by electrostatic or electromagnetic means. In the electrostatic type, the beam is deflected by an electric field set up across the deflection plates by a deflection voltage. In the electromagnetic type the beam is deflected by a magnetic field established by a deflection current in a coil around the outside of the tube.

A simplified arrangement of the construction details of a cathode-ray tube employing electrostatic deflection and focusing is shown in figure 13-1. The cathode when heated

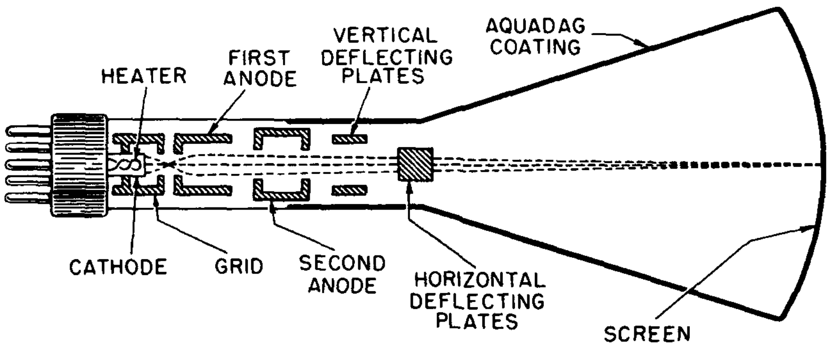


Figure 13-1.—Construction of cathode-ray tube using electrostatic deflection and focusing.

by the enclosed filament releases free electrons. A cylindrical grid surrounds the cathode and controls the beam intensity as electrons pass through the end-opening of the grid. The control is accomplished by varying the negative voltage on the grid and is called INTENSITY OR BRIGHTNESS CONTROL. After leaving the grid the electron stream passes through two or more cylindrical anode focusing plates which concentrate the electrons into a narrow beam. The first anode concentrates the free electrons and the second anode accelerates them. The entire assembly including the cathode, grid, and the two anodes is called the ELECTRON GUN. The electrons emerge from the electron gun at high speed.

The grid helps to narrow the beam but cannot focus it to a sharp point on the viewing screen. The two anodes aid in the focusing action, as shown in figure 13–2. The electric

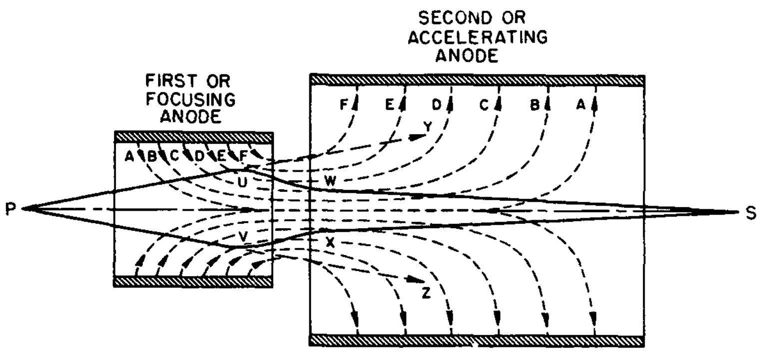


Figure 13–2.—Electrostatic focusing.

field is established between the anodes. Electrons entering this field converge at point S on the screen. The second anode is positive with respect to the first anode and both are positive with respect to the cathode in order to attract electrons from the cathode and to accelerate them. The repelling force of like charges tends to scatter the electrons but they are accelerated to such a high speed that the scattering action is not effective in defocusing the beam. Never-

theless the mutual repulsion between electrons determines the sharpness with which a beam may be focused on the screen.

The focus of the electrostatic type of cathode-ray tube is generally controlled by varying the voltage between the first anode and the cathode. This voltage varies the force exerted on the electrons and tends to narrow the beam. Thus if the screen is observed when the first anode voltage is varied, the beam may be brought to a bright sharp spot.

Focusing in an electromagnetic cathode-ray tube is accomplished by a coil encircling the outside neck of the tube. The coil may be moved along the neck to a limited extent to focus the beam but the normal method, after the coil is in the proper position, is to vary the current flowing through the coil.

Without lateral deflection, the electron gun produces only a small spot of light on the viewing screen. With deflection, the trace of the spot forms a line on the screen. The electrostatic type cathode-ray tube uses two pairs of deflection plates mounted at right angles with respect to each other, as shown in *figure 13–3*. The *vertical deflection* plates

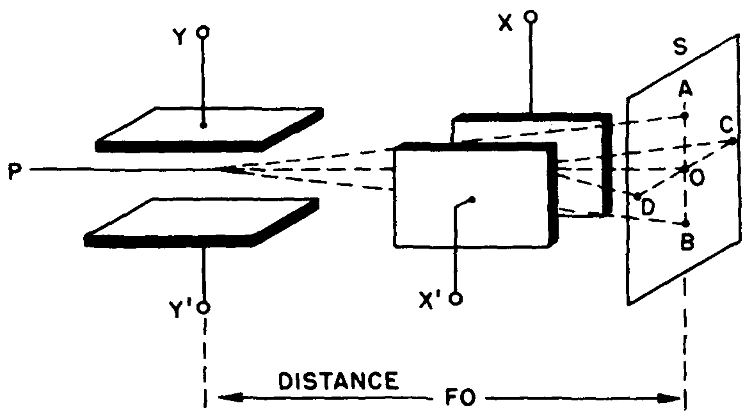


Figure 13–3.—Deflecting plates for electrostatic cathode-ray tube.

(YY') deflect the beam in a vertical direction, and the horizontal deflection plates (XX') deflect it horizontally. Both pairs usually function simultaneously. The beam is attracted by the positive plate and repelled by the negative plate as the electrons pass between them. One plate of each pair is usually grounded. To deflect the beam, a positive or negative voltage is applied between the other plate and ground, thus establishing an electric field between the plates. The deflecting force varies with the deflection voltage across the plates and with the field intensity.

If plate Y is positive with respect to Y' the beam is deflected upward, striking the screen at A. If plate Y is negative with respect to Y' the beam is deflected downward, striking the screen at B. If there is no deflection voltage across the plates the beam will strike the screen at O. The amount of deflection varies with the deflection voltage across the plates. If plate X is positive with respect to X' the beam will be deflected horizontally and strike the screen at C. If X is negative with respect to X' the beam will strike the screen at D. Both pairs of plates are mounted near the output end of the electron gun with the vertical deflection plates farthest from the screen. Centering controls are provided, which enable the operator to move the spot to any desired point on the screen.

The electron beam deflection angle is the angle through which the beam may be deflected in any direction from the center line through the tube. A cathode-ray tube having a 50° deflection angle is one in which the electron beam can be deflected in any direction at an angle of 25° with respect to the center line. Such tubes are called WIDE ANGLE TUBES and are constructed with a greater flare and shorter length so that the beam can be deflected through a large angle without striking the tube walls. Wide-angle cathode-ray tubes are used extensively in television receivers.

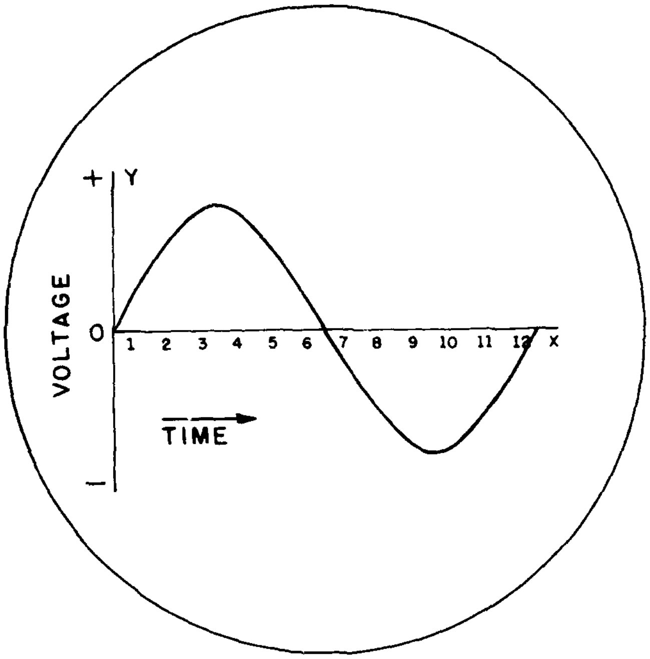
The length of time that the screen glows or fluoresces at the point where the electron beam strikes it depends on the material of the phosphor coating on the screen, and is known as SCREEN PERSISTENCE. Some cathode-ray tubes have a

long persistence screen and others a short persistence, depending on their use. The screen phosphors are designated by the letter "P" followed by a number. Most radar indicator cathode-ray tubes employ $P1$, $P4$, or $P7$ screen phosphors. The $P1$ and $P4$ phosphors have medium persistence and give off green and white light respectively. The $P7$ screen has long persistence and gives off yellow light.

All fluorescent materials have some phosphorescence, or afterglow, but the duration of the afterglow varies with the material, as well as with the amount of energy in the beam causing the emission of light. For oscilloscopes that are to be used for observing nonrepeating phenomena or periodic phenomena that occur at a low repetition ratio, a screen material on which the image will linger is desirable. In applications where the image changes rapidly, prolonged afterglow is a disadvantage, because it may cause confusion on the screen.

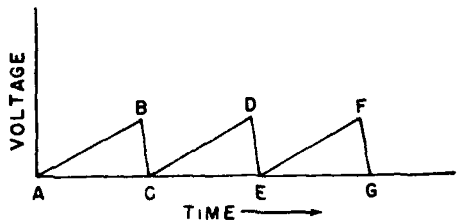
The eye retains an image for about one-sixteenth of a second. Thus in a motion picture the illusion of motion is created by a series of still pictures flashed on the screen so rapidly that the eye cannot follow them as separate pictures. In the cathode-ray tube the beam is repeatedly swept across the screen and the series of adjacent spots appears as a continuous line. Thus the wave shape of an a-c voltage can be observed on the screen when the a-c voltage is applied to one pair of deflection plates and simultaneously a second voltage of appropriate characteristics is applied to the other pair of plates.

The conventional way of representing voltage or current of sine waveform is shown in figure 13-4, A. The voltage to be observed is applied across the vertical deflection plates and simultaneously a saw-tooth voltage is applied across the horizontal deflection plates. The saw-tooth voltage moves the beam from left to right at constant speed to form the time scale along OX; then it returns the beam rapidly to the starting position at the left and repeats the operation. The saw-tooth voltage is so named because when plotted against time it resembles a saw-tooth as shown in figure 13-4, B.



A

SINE-WAVE VOLTAGE PLOTTED AGAINST TIME



B

SAW-TOOTH WAVEFORM PLOTTED AGAINST TIME

Figure 13-4.—Sine-wave and saw-tooth voltage waveforms.

As the voltage increases from A to B the beam is swept from 0 to 12 (fig. 13–4, A). As the voltage falls from B to C in figure 13–4, B, the beam is quickly returned to its starting position and the process is repeated.

If an a-c voltage of sine waveform is placed across the vertical deflection plates with no horizontal deflection, a single vertical line appears on the screen. The varying rate of change of the voltage is hidden because the vertical movements retrace themselves repeatedly on the same vertical line. Similarly, if a sweep voltage of saw-tooth waveform is applied to the horizontal deflection plates in the absence of vertical deflection, a horizontal line is formed and the rate of change of the voltage is obscured. However, when both voltages are introduced at the same time, the vertical motion of the beam is spread out across the screen to form a sine curve like that shown in figure 13–4, A.

Oscilloscope Circuits

A block diagram of a cathode-ray oscilloscope is shown in figure 13–5. The horizontal deflection amplifier is a high-gain R-C coupled class-A wide-band voltage amplifier

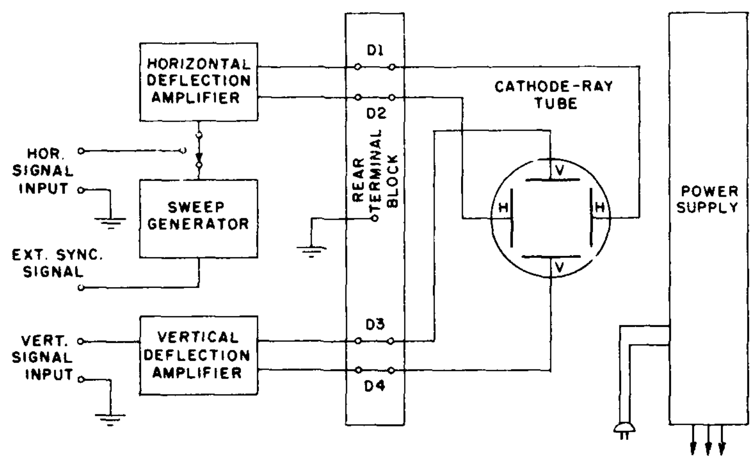


Figure 13–5.—Block diagram of a cathode-ray oscilloscope.

that increases the amplitude of the horizontal input voltage and applies it to the horizontal deflection plates. The sweep generator supplies a saw-tooth voltage to the input of the horizontal amplifier through a switch that provides an optional external connection. The vertical deflection amplifier increases the amplitude of the vertical input voltage before applying it to the vertical deflection plates. The input to the vertical amplifier appears in magnified form on the viewing screen as a graph of the current or voltage waveform being examined. A rear terminal block provides direct electrical connections to the deflection plates. The direct connections are used, for example, when examining d-c potentials, or high-frequency signals that would be attenuated excessively by the amplifier circuits. The power supply provides all d-c voltages for the tubes including a high d-c potential for the cathode-ray tube.

A schematic diagram of an elementary cathode-ray oscilloscope is shown in figure 13–6. The cathode-ray tube employs electrostatic focusing and deflection. *V*1 is the vertical amplifier, *V*2 the horizontal amplifier, and *V*3 the sweep generator. *R*1 is a manual vertical gain control, *R*2 is a manual horizontal gain control, *S*3 is the coarse frequency adjustment for *V*3, and *R*10 is the fine frequency adjustment. An external synchronizing signal may be applied from an external source to the grid of *V*3 when *S*2 is in the EXT. SYNC. position. The synchronizing signal is obtained from the plate of *V*1 when *S*2 is in the INT. SYNC. position. *R*3 provides manual control of the sync signal amplitude.

The low voltage power supply includes a conventional full wave rectifier, *V*4; secondary windings *L*4, *L*5, and *L*6 of the power supply transformer; and the pi-filter (*C*17, *L*13, and *C*18). The cathode of *V*4 is positive with respect to ground. The output voltage (440 volts) is applied to the plate circuits of *V*1, *V*2, and *V*3. The voltage divider (*R*14, *R*15, and *R*19) is connected across the output of *V*4 and supplies +170 volts to the left side of the centering controls (*R*17 and *R*18).

The high voltage power supply includes *V*5 and secondary

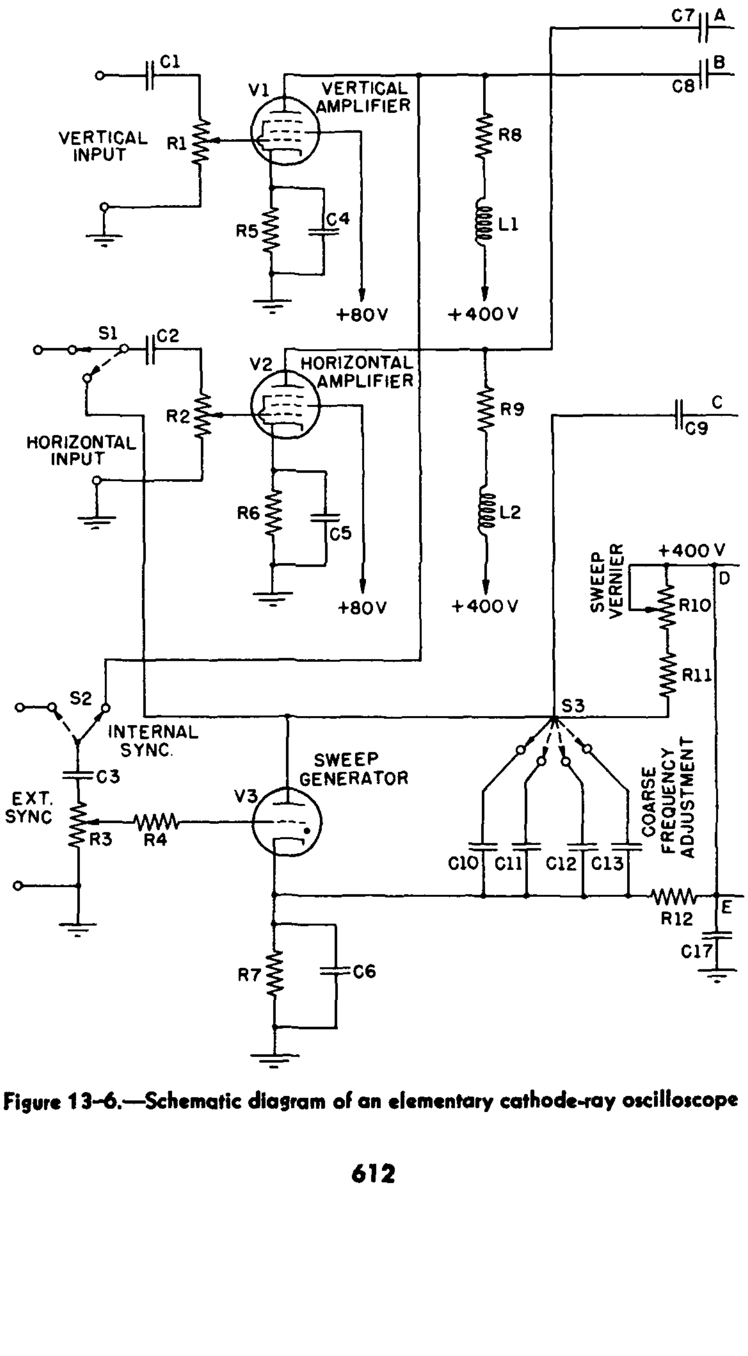


Figure 13-6.—Schematic diagram of an elementary cathode-ray oscilloscope

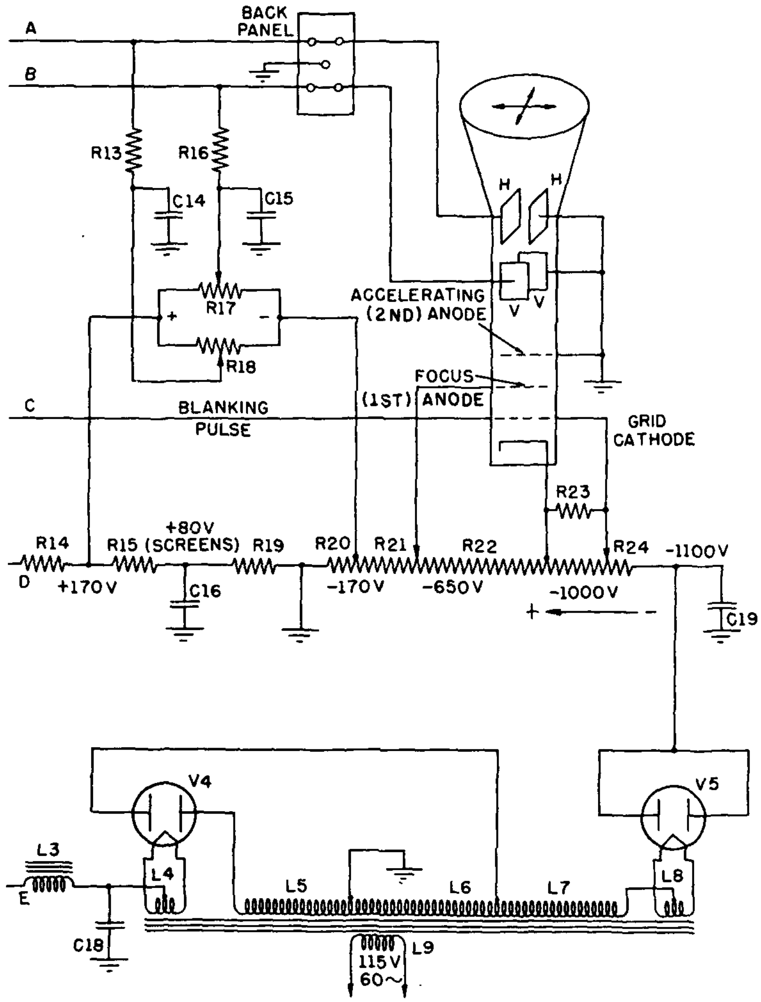


Figure 13-6.—Schematic diagram of an elementary cathode-ray oscilloscope—Continued.

windings $L6$, $L7$, and $L8$ of the power supply transformer. The output voltage of $V5$ is negative with respect to ground and is applied across the voltage divider ($R20$, $R21$, $R22$, and $R24$). $C19$ filters the output voltage. The tap between $R20$ and $R21$ supplies −170 volts to the right side of $R17$ and $R18$.

The cathode of the cathode-ray tube is connected at −1,000 volts with respect to the ground. The second anode in the cathode-ray tube is grounded, and the first anode is negative with respect to the second anode but both anodes are positive with respect to the cathode. This arrangement provides the necessary accelerating voltage for the electrons in the beam to form a bright spot on the screen and at the same time prevents defocusing the spot by holding the average voltage across the deflection plates close to the potential of the second anode. The arrangement also introduces a safety factor by removing high voltage from the deflection plates and the associated input terminals on the rear panel.

Capacitors $C1$, $C2$, and $C3$ block any external d-c voltage components from the grids of $V1$, $V2$, and $V3$. Similarly, capacitors $C7$ and $C8$ block the d-c components of plate voltage from the cathode-ray tube deflection plates and at the same time couple the a-c components to them. $C9$ couples a blanking pulse to the grid of the cathode-ray tube, which blanks out the return trace of the sweep generator. During the time the sweep voltage rises in a positive direction, $C9$ charges at a constant rate through $R24$ and the *C-R* tube bias is reduced accordingly. As the sweep voltage suddenly falls and snaps the electron beam back to the left side of the screen, $C9$ rapidly discharges through $R23$, driving the cathode more positive, and biases the *C-R* tube below cutoff so that the return trace is invisible.

The synchronizing voltage applied to the grid of $V3$ stabilizes the screen pattern, as described in chapter 7.

$R17$ and $R18$ are positioning controls that provide manual adjustment of the low d-c voltages that may be applied across the two pairs of deflection plates. The spot is approximately centered on the screen when the voltage between the

contact arms of $R17$ and $R18$ is zero. Moving the contact arm of $R17$ to the right makes one vertical plate more negative and thus repels the beam and moves the spot vertically a certain distance on the screen. Conversely moving the contact arm of $R17$ to the left of the zero position makes the vertical plate more positive, and thus attracts the beam and moves the spot in the opposite direction.

Appliactions

Observation of waveforms.—The cathode-ray oscilloscope is generally used to observe voltage waveforms in testing electrical circuits. The electrostatic cathode-ray tube employs voltage sources rather than current, to deflect the electron beam. For this reason the electrostatic type of cathode-ray tube is used in test oscilloscopes. The electromagnetic cathode-ray tube is a current-operated device. It is used in certain applications other than general testing, where its properties make it more suitable than the electrostatic tube.

To obtain an accurate representation of the voltage waveform, a few precautions must be observed. For the protection of both the operator and the oscilloscope, the approximate magnitude of the voltages in the circuit under test must be known. Dependable data can be obtained from the oscilloscope only if its sensitivity and its frequency characteristics are known. To make certain that the waveform will not be distorted, it is essential that the manner in which distortion takes place be understood and that precautions be taken to minimize such distortion.

The input to most oscilloscopes is between an input terminal and ground. The input terminal is coupled to the amplifier grid through a capacitor whose voltage rating rarely exceeds 450 volts. Therefore, unless the approximate magnitude of the voltage under test is known, damage to the oscilloscope through breakdown of the input capacitor may occur.

In some cases it may be necessary to observe waveforms

in circuits where the voltage is much greater than that which the components within the oscilloscope can withstand. A voltage divider may be used in such instances to reduce the voltage to a value that will not damage the equipment. In any case it is important that the oscilloscope be adequately grounded—a precaution that must be taken for the protection of the operator, because a failure of some part of the voltage divider can raise the potential of the whole oscilloscope to a dangerous level if the case is not solidly connected to ground.

If a capacitance voltage divider is used, a wise precaution is to shunt each capacitor with a high resistance to maintain the proper voltage distribution across each capacitor.

The range of sweep frequencies in a given oscilloscope is usually indicated on the control panel, as shown in figure 13–7. The sweep frequency generator in this example has a frequency range of 3 to 50,000 cps. The frequency range that the vertical and horizontal amplifiers are capable of amplifying properly is given in the applicable manufacturer's instruction book. In this example the vertical amplifier has a bandwidth of from 30 cps to 2 mc and the horizontal amplifier has a bandwidth of from 25 cps to 100,000 cps. Generally, only the best oscilloscopes use amplifiers that will amplify voltages whose frequency is below 30 cps or above 100,000 cps. Oscilloscopes that do not cover as wide a range of frequencies as the one shown in figure 13–7 may be satisfactory for most uses, but distortion is likely to occur when saw-tooth or rectangular waveforms of a high recurrence rate are investigated.

The deflection sensitivity may be expressed as the distance in millimeters that the spot is moved on the screen when 1 volt is applied across one pair of deflection plates. The deflection sensitivity of the vertical deflection plates in the oscilloscope shown in figure 13–7 is 0.528 millimeters per volt. Expressed in volts per inch of deflection, the deflection becomes $\frac{25.4}{0.528}$, or 48 volts per inch. The deflection sensi-

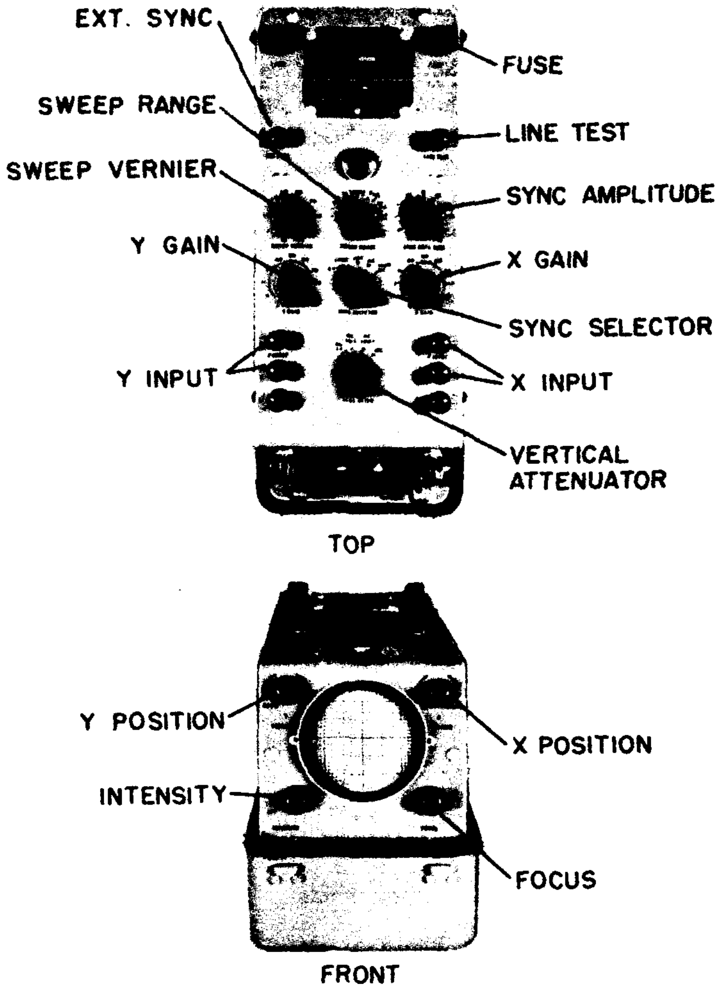


Figure 13-7.—Oscilloscope controls.

tivity of the horizontal plates in this example is 0.379 millimeters per volt, or 67 volts per inch.

The deflection sensitivity may also be expressed as the input voltage to the amplifier (horizontal or vertical) for a deflection of 1 inch of the spot on the screen. In this case

the amplifier gain control is adjusted to a suitable value that is arbitrary (for example, mid scale). In the example of figure 13–7, both the horizontal and vertical deflection sensitivity are 0.1 volt rms for 1 inch peak-to-peak deflection.

To avoid pick-up of stray signals the leads from the circuit under test should be as short as possible, and they should be shielded. The cathode-ray tube itself is shielded by the aquadag coating on the inside of the tube and a metal shield on the outside. A common side of the oscilloscope circuits is grounded and should be connected to a ground point in the circuit under test and to a good external ground connection.

Several causes of distortion are possible in the production of a cathode-ray tube display. Some of these causes are:

1. Exceeding the bandwidth limitation of the deflection amplifiers.
2. A defective sweep generator.
3. Excessive fly-back time.
4. Excessive synchronizing voltage.
5. Oscilloscope loading of a high impedance test circuit.
6. Oscilloscope capacitance shunting of video amplifier test circuit.
7. Use of a variety of oscilloscopes and test leads on one equipment.
8. Improper shielding of test leads.

SYNCHROSCOPE

Components

An oscilloscope that has a sweep of very short duration generated only when a synchronizing signal is provided is called a SYNCHROSCOPE. A block diagram of a simplified synchroscope is shown in figure 13–8. A signal pulse triggers the sweep circuit, which has a saw-tooth waveform in which the time of one sweep is small compared with the time between successive sweeps. An alternate method of triggering the sweep is by a separate timing generator (not shown in the figure). In either case the sweep is initiated

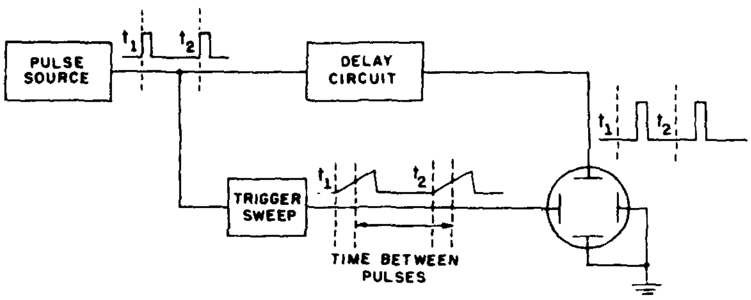


Figure 13-8.—Block diagram of a simplified synchroscope.

before the signal pulse reaches the deflection plates. The delay circuit permits the sweep circuit to be initiated before the application of the pulse to the vertical plates. This action causes the entire pulse including the leading edge to appear on the screen.

Synchroscopes have several calibrated sweeps—for example, the sweep may be 50 microseconds, 200 microseconds, or 1,000 microseconds in length. Such instruments are useful in observing the waveform of very short pulses like those from a radar equipment, or in observing the time interval between pulses and the duration of a pulse. A very important function of a synchroscope is to observe the standing-wave ratio in a waveguide between a radar transmitter and antenna in tuning and checking the equipment for optimum performance.

ELECTRONIC SWITCHING

An electronic switch is a device that utilizes the properties of gas-filled or high-vacuum tubes for closing, opening, or changing the operation of an electronic circuit. The electronic switch is more sensitive than a mechanical switch. It is very fast and is usually silent in operation. For example, an electronic switch may be used to GATE an amplifier circuit—that is, to cause it to function during a given period of time and to prevent it from functioning at other times.

The on-and-off periods of operation may be the same or they may be different, depending on the operating requirements.

Multivibrator Used as an Electronic Switch

A multivibrator may be used as in figure 13-9, A, to cut an amplifier tube, *V*1, on and off at the multivibrator frequency. The waveforms shown in figure 13-9, B, indicate the manner in which *V*1 is gated. The multivibrator output, e_k, composed of essentially square waves, is developed cross the cathode resistor of *V*3. *V*3 conducts periodically and develops a positive gate voltage, e_k, between the cathode and ground of *V*1. The gate voltage cuts off *V1* during the time *V*3 conducts. If the input to *V*1 is a series of regularly spaced positive voltage spikes, as shown in figure 13-9, B, the gate voltage permits pulses to pass through *V*1 and blocks off two pulses. If the gate voltage existed for a longer period of time, three or more pulses could be passed and an equal number suppressed.

The switching action of a multivibrator may also be used to show, for purposes of comparison, two or more signals on a cathode-ray oscilloscope apparently at the same time. Although the signals are not actually present at the same time they appear to be so because of persistence of vision of the human eye.

An arrangement for presenting two signals at the same time is shown in figure 13-10. Tubes *V*1 and *V*2 with their associated circuits make up a conventional multivibrator. *V*3 and *V*4 separate amplifier stages having a separate input and a common output via *C*5 to the vertical deflection plates of an oscilloscope.

The circuit operates as follows: So that the signal will be stationary on the oscilloscope, assume that input signals *A* and *B* have the same frequency and that input signal *A* is used to synchronize the multivibrator. When *V*1 conducts, *V*3 is cut off and the amplified signal from *V*4 appears on the oscilloscope. When *V*2 conducts, *V*4 is cut off and the amplified signal from *V*3 appears on the oscilloscope. It is assumed that the time of one positive pulse from the multi-

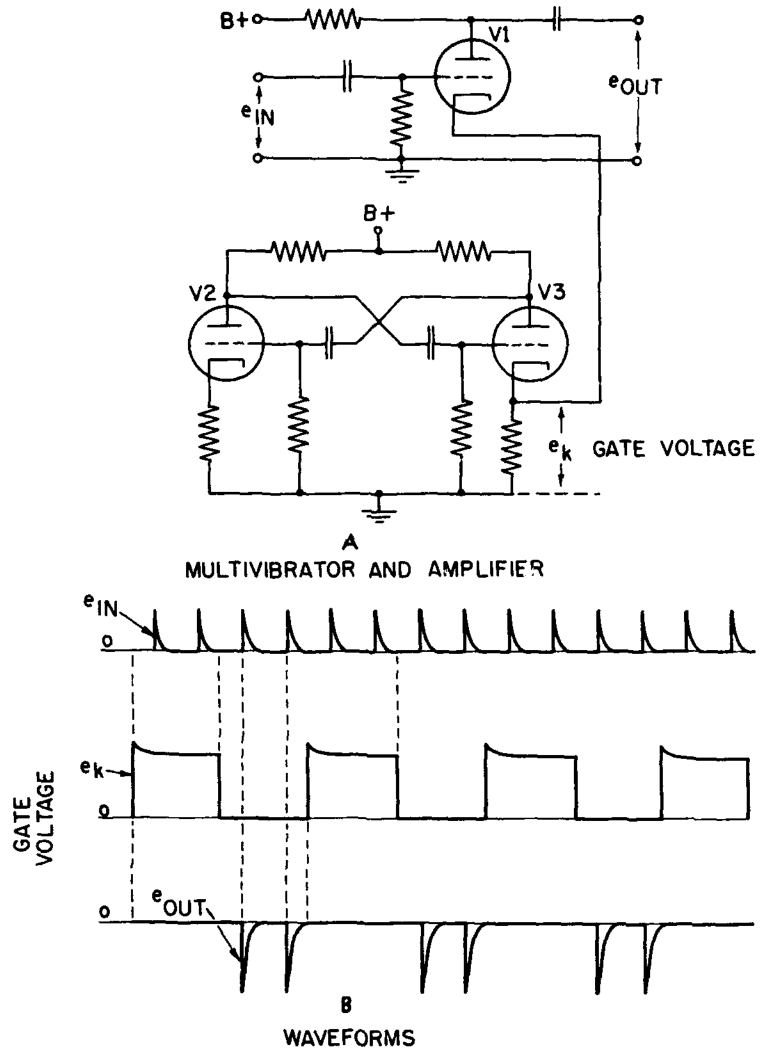


Figure 13-9.—Multivibrator electronic switch used to provide a gate voltage, and waveforms.

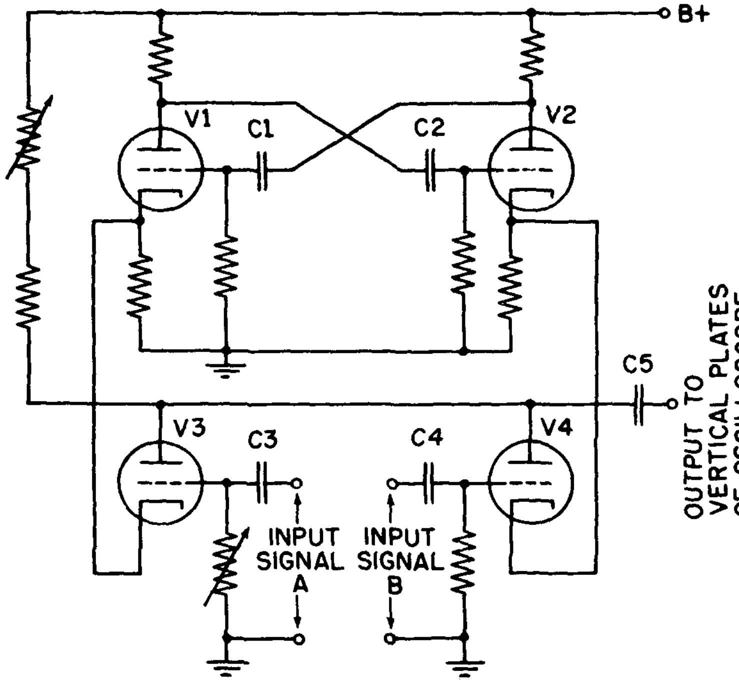


Figure 13-10.—Multivibrator electronic switch for use with a cathode-ray oscilloscope.

vibrator is equal to that of one or more periods of input signal *A*.

By connecting the output of the electronic switch to the vertical amplifier on the oscilloscope and by proper adjustment of the level controls on the electronic switch (not shown in the figure) the two signals are made to appear on the same horizontal axis on the oscilloscope screen. By adjusting the external synchronizing signal control on the oscilloscope the two signals are made to start at the same point on the horizontal axis. The external synchronizing signal may originate with either signal *A* or signal *B*.

ABSORPTION WAVE METER

Absorption wave meters are used to measure the frequency of the higher r-f sources. They operate on the principle that the current in an *L-C-R* circuit, coupled to the source being measured, will increase to a maximum when the *L-C-R* circuit is tuned to the resonant frequency of the source. Thus at relatively low frequencies (up to about 30 mc) a simple tuned circuit like the one shown in figure 13–11 may be employed. The inductor, *L*, is usually wound on a standard plug-in coil form so that a number of coils, each covering a specific range of frequencies, may be used. The capacitor,

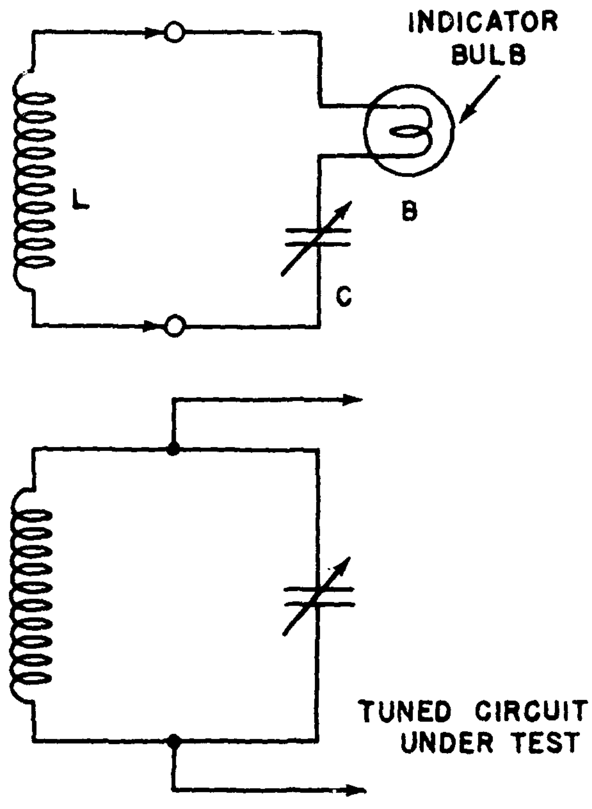


Figure 13–11.—Absorption wave meter.

C, is variable so that the circuit may be brought into resonance at a particular frequency in any of the ranges covered by the plug-in coils. The capacitor is calibrated and with this calibration the resonant frequency can be readily determined.

The current indicator may be an r-f milliammeter, or in less expensive types, a low-voltage incandescent flashlight bulb (B in the figure). Since absorption wave meters tend to detune self-excited oscillator circuits to which they are coupled, they cannot be relied on for very accurate measurements. Because such meters are relatively insensitive, they indicate only the fundamental frequency. At high frequencies the resonant circuit may be Lecher wires, coaxial stubs, butterfly circuits, or tuned cavities.

GRID-DIP METER

The grid-dip meter is a versatile and popular test instrument. A simplified circuit of the grid-dip meter is shown in figure 13–12.h Wit this meter it is possible, among other

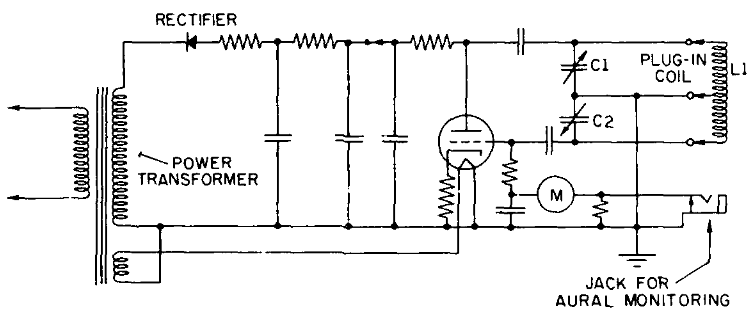


Figure 13–12.—Schematic diagram of a grid-dip meter.

things, to determine the resonant frequency of an antenna system, to detect harmonics, and to check relative field strengths—for example, in rotating a television antenna for maximum signal strength. The meter may also be used as an absorption frequency meter when the oscillator is not energized.

Basically this instrument is a calibrated oscillator in which provision is made for metering the grid current in the oscillator circuit. With the oscillator functioning, energy is coupled from the tuned circuit (composed of $C1$, $C2$, and $L1$) to the circuit under test. The circuit under test is supplied a small amount of energy via tank coil $L1$ of the meter. Except for the field of $L1$, the circuit under test is deenergized. The capacitors are rotated to the point where the oscillator tank frequency is equal to the resonant frequency of the circuit under test. At resonance the grid current decreases as indicated by the dip in the grid meter. The energy absorbed from $L1$ by the circuit under test decreases the a-c component of plate voltage thus causing a decrease in feedback energy from the plate to the oscillator grid. The grid voltage is driven less positive and the grid current decreases.

In order to determine the resonant frequency of an antenna system, coil $L1$ of the meter is brought into close proximity to the antenna when the latter is deenergized. The proper point along the antenna is the point corresponding to a high-current point when the antenna is energized. If the antenna has its center open, a jumper is used as a temporary short so that the test may be made. The meter is tuned until the dip of the grid current indicates resonance. Harmonics may also be indicated in the same manner.

Standing waves on an open transmission line may be checked by removing the plate power supply and operating the tube as a diode detector. This action is similar to that of an absorption wave meter.

FREQUENCY STANDARDS

Primary Frequency Standard

When a major naval operation is planned, surface ships, submarines, carrier-based planes, and other units may be required to maintain radio silence until contact with the enemy is made. Separate movements are then synchronized

by means of preset-communication frequencies. One of the most important assignments of a technician is to have radio receivers and transmitters accurately adjusted to the frequencies assigned by the task-force communications officer. Such an assignment involves the use of a frequency standard of high accuracy, such as the LR or LM frequency meter. If the frequency meter is out of order (for example, as the result of a defective reference crystal), it becomes necessary to replace the crystal and verify the accuracy of the new reference crystal. The primary frequency standard is station WWV operated by the National Bureau of Standards. Every frequency meter should be checked weekly against the WWV transmissions to assure that the reference crystal is still reliable. Standard-frequency transmissions are useful also as standards for measuring field intensity and audio frequencies.

The schedule of services offered by station WWV is published in Radio Navigational Aids, H. O. 205, issued by the USN Hydrographic Office. For current information refer to this publication. Revisions to the schedule of services are reported from time to time in nearly all naval electronic magazines.

Secondary Frequency Standards

The LM frequency meter is a secondary frequency standard of high accuracy. The simplified block diagram shown in figure 13–13 shows the basic components of the LM–18 meter and indicates how they function. This meter is fundamentally a stable, self-excited heterodyne oscillator of the electron-coupled type covering the range from 125 kc to 20 mc and having a separate crystal-controlled reference oscillator incorporated within it. Zero beats are provided at several reference points between the two oscillators.

As indicated in figure 13–13, A, the crystal oscillator serves as a means of checking the heterodyne oscillator frequency. A small trimmer capacitor is connected in parallel with the main tuning capacitor and serves to correct the frequency of the heterodyne oscillator at the nearest crystal check point.

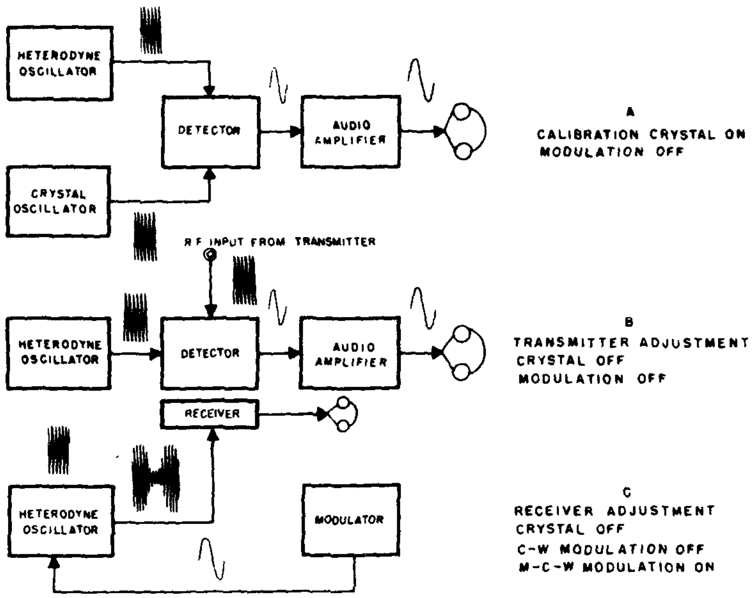


Figure 13-13.—Simplified block diagram of the LM-18 meter.

The beat frequency between the two oscillators is detected, amplified, and fed to the headphones for aural indication. At zero beat, the heterodyne oscillator frequency is correct for the dial setting as indicated in the calibration book, for the selected crystal check point.

For transmitter adjustment, the heterodyne oscillator, after calibration, is combined with the r-f signal input from the transmitter, as shown in figure 13-13, B. Zero beat results when the transmitter is adjusted to the frequency of the oscillator. Aural indication is accomplished in the same manner as calibration of the heterodyne oscillator.

The equipment may serve as a signal source for alignment and calibration of receivers, as indicated in figure 13-13, C. After calibration to the nearest crystal check point, the output of the heterodyne oscillator is fed to the receiver. An r-f attenuator is provided for adjustment of the output signal

level. The receiver is tuned to zero beat with the heterodyne oscillator, as indicated by means of the headphones in the output circuit of the receiver. By means of circuit switching the audio amplifier serves as a 500-cycle modulator. For a-m input signals, receiver calibration is performed by tuning for maximum audio output.

The output of the LM meter is too low for some maintenance uses for which less accurate signal generators with higher output power are provided. Since the oscillator may be set within 1 part in 10,000 for the range from 2 to 20 mc, careful use of the LM meter should result in close agreement of all transmitters and receivers set to any one frequency.

Other secondary frequency standards used by the Navy include the LR and TS–186/UP. The LR covers 15 ranges from 160 kc to 30 mc, with harmonic extensions on some frequencies above 30 mc. The accuracy of the r-f output of the LR is within 0.003 percent. The TS–186/UP covers the range from 100 to 10,000 mc having an over-all accuracy of 0.01 percent.

R-F Signal Generator

The output power of a secondary frequency standard or frequency meter is too low to energize directly an output meter when radio equipment is being serviced. Signal generators, such as Navy types LP and LX, have sufficient power for general service work. Some signal generators include vacuum-tube voltmeter circuits, modulating circuits, and others. The various test sets (for example, TS–382/U and TS–535/U) employ auxiliary circuits for specialized test work.

Basically an r-f signal generator is a device used for the production of r-f voltages by means of a well-shielded oscillator. It has a frequency-calibrated dial, a provision for adjusting the amplitude of the output voltage, and a means for modulating the r-f signal if desired. Other refinements may also be included.

By the application of a proper r-f signal it provides a means of aligning the r-f and i-f circuits of a receiver. If the equip-

ment is sufficiently accurate and if a detector circuit together with one or more audio stages is included, the device may be used to determine an unknown frequency by the zero-beat method.

To afford a better understanding of the construction and function of an r-f signal generator, the Navy model LAH signal generator is described briefly. A simplified circuit diagram of this meter is shown in figure 13–14.

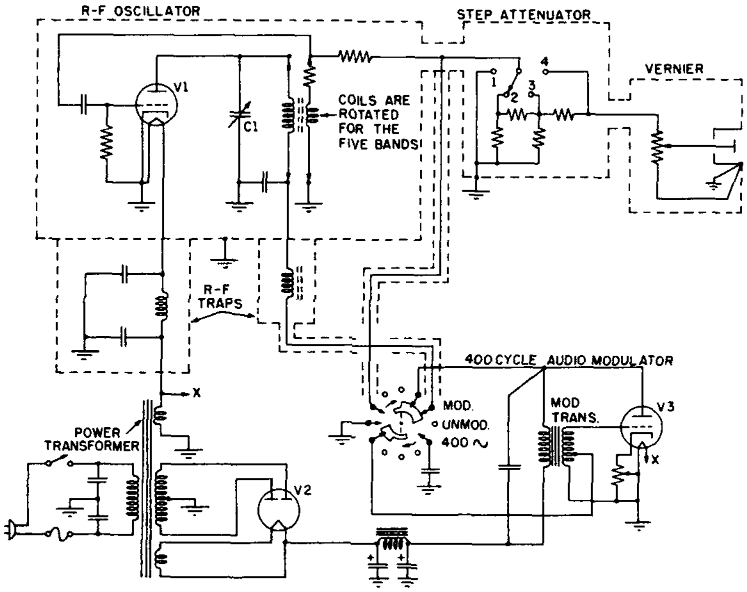


Figure 13–14.—Simplified diagram of Navy model LAH signal generator.

The r-f oscillator tube, $V1$, is operated as a tuned-plate oscillator. Five sets of coils (not shown in the figure) in conjunction with $C1$ permit the set to operate on five frequency bands covering the range from 100 kc to 32 mc, all operating on fundamental frequencies. One r-f trap in the filament circuit of $V1$ tends to keep r-f out of the filament. Another trap in the plate circuit of $V1$ keeps r-f out of the modulation transformer.

The *V*3 and the modulation transformer constitute the 400-cycle internal audio oscillator, which provides the necessary energy to modulate the r-f oscillator at approximately 30 percent.

The attenuator is built in two sections. The first section may be adjusted to give four levels of attenuation (approx. 10-to-1 per step). The second section permits a vernier adjustment of the output. The attenuator is so designed that it presents a constant load to the oscillator.

This type of oscillator may be used with an output meter for the alignment of radio sets.

A-F Signal Generator

A signal generator that would incorporate all types of signals required in the servicing of electronic equipment would be too expensive and bulky for practical use. Therefore, the present trend in test equipment is toward signal generators that can be used individually or in combination with auxiliary equipment. The a-f signal generator is no exception.

A-f generators are used in plotting response curves on audio amplifiers, transformers, and other audio circuits. They are also used in plotting curves on filter circuits, for supplying audio modulation to r-f signal generators, and for numerous other applications.

One common type of audio oscillator is the beat frequency oscillator (BFO) of which Navy model LO–3 is an example. The block diagram of this model is shown in figure 13–15.

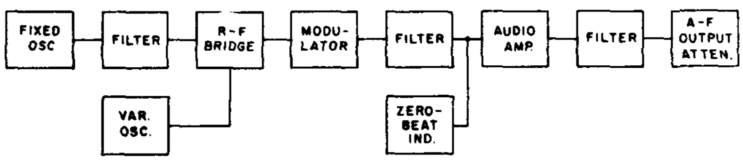


Figure 13–15.—Block diagram of Navy model LO–3 beat-frequency oscillator.

It is composed of a fixed frequency oscillator that operates at a frequency of 250 kc, and a variable frequency oscillator that operates in the range between 235 kc and 250 kc. The resulting beat extends from zero to 15 kc.

Two oscillators supplying a common load and operating at nearly the same frequency will tend to lock into synchronism with each other and thus produce a zero beat. The r-f bridge isolates the two oscillators and prevents them from pulling into step as they approach the same frequency (at low a-f settings on the dial). The bridge circuit permits a more accurate calibration at the low frequencies. The filter following the fixed oscillator prevents harmonics in the fixed oscillator from entering the modulator. The filter following the modulator blocks r-f from the audio-amplifier grid. The filter following the a-f amplifier completes the filtering and prevents any remaining r-f currents from reaching the output.

To make the frequency calibration accurate under conditions of changing ambient temperature, aging of tubes and components, and so forth, a zero-beat indicator is included in the circuit. The input of the indicator tube is connected to the output of the modulator and as the frequency dial is turned toward zero the slow beating of the oscillators is observed in the indicator tube.

The output of the oscillator is controlled by a constant-impedance pad.

RADIO-INTERFERENCE FIELD-INTENSITY METER

Field-strength meters covering the various frequency ranges have been developed for locating r-f interference. An example of an h-f and v-h-f radio-interference field-intensity meter is the TS–587/U.

A field intensity meter is essentially a portable radio receiver with an attached meter to indicate the strength of the received signal. It is useful in locating the source of an interfering signal on own ship; for testing the effectiveness of measures for eliminating interference; for seeking out the sources of radiation on own ship that violate radio silence; and for plotting field patterns of directive antenna arrays.

It is less sensitive than a radio-countermeasures receiver, which may also be used to detect unauthorized transmission but which is handicapped by lack of mobility. Direction finders may also be used for this purpose.

Figure 13–16 illustrates one use of the OF interference

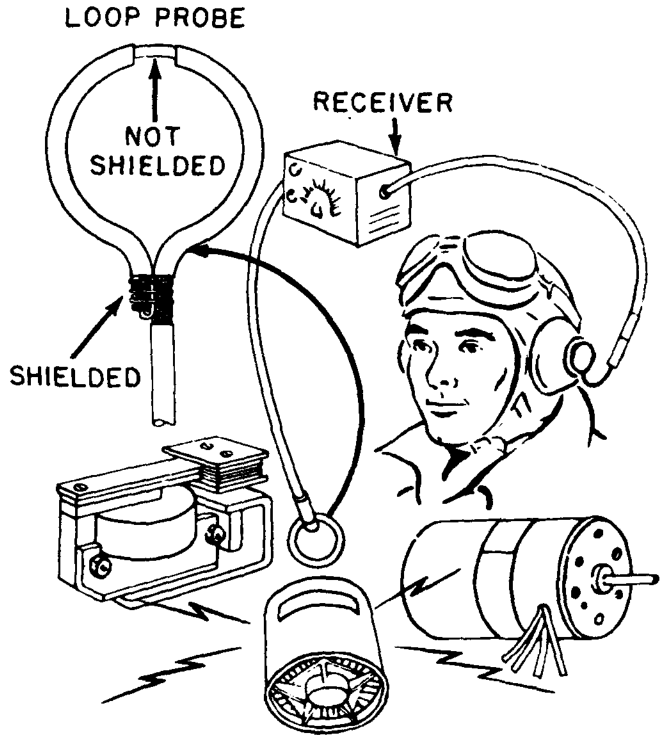


Figure 13–16.—Location of radio noise.

locator, and figure 13–17 is a block diagram of this equipment. The OF equipment is an early type of field strength meter and is representative of the class. Other types, such as the TS–576/U, have been developed that have wider ranges and greater sensitivity.

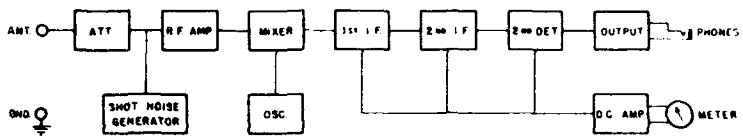


Figure 13–17.—Block diagram of model OF–2 interference locating equipment.

The block diagram of the OF–2 (fig. 13–17) indicates it to be essentially a superheterodyne radio receiver. The circuit is conventional except for the addition of the noise generator and the meter circuits.

Control-grid bias for the first and second i-f tubes (having remote cutoff characteristics) is supplied by the a-v-c system. This feature enables the voltage developed across the diode load resistor to be stabilized at a value approximately proportional to the logarithm of the input signal. This action is necessary in order to provide an indicating meter scale with a range extending from 10 to 1,000 microvolts and with uniform reading accuracy over the entire range.

A diode is used to supply the a-v-c voltage, and a portion of this voltage is fed to the d-c amplifier which has the indicating meter in the plate circuit. Thus, while the meter reads the rectified a-v-c voltage, it is effectively reading signal voltage because, as mentioned previously, the a-v-c voltage is proportional to the logarithm of the signal voltage. A battery is provided in the plate circuit of the diode to buck out the indicating meter current under zero input conditions In this way the meter is set for zero under conditions of no-signal input. Another diode rectifies the signal to provide an audio output for the headphones.

The shot-noise generator is employed for calibration purposes. The output of the generator is set, in the process of calibration, to duplicate the strength of the signal under test. Because the output of the shot-noise generator is known, it is thus possible to measure the unknown signal strength by comparison.

Because power for the set is supplied entirely by dry

batteries, no rectifiers or power transformers are necessary, and the equipment is portable.

Measurement and Location of Interference

To locate a source of radio interference, the interfering signal is tuned in on the radio-interference field-intensity meter. Earphones should be used to identify the signal being received, as in figure 13–16. An electrostatically shielded loop probe (shown enlarged in fig. 13–16) may be used as an antenna to locate source of noise in machinery. Moving the probe in the direction of the source (or some conductor radiating the noise energy) causes the signal strength, as indicated by the meter or earphones, to increase. Moving the probe away from the source causes the signal strength to decrease. Inspection of all rotating equipment usually is necessary to locate interference on shipboard, and a final check should be made by starting and stopping the suspected device.

Occasionally v-h-f transmitters indirectly cause interference in lower frequency bands. For example, on one carrier a search radar was causing interference with all TBS and other radio reception. Two steel-wire stays, in loose contact near the radar antenna, were causing the interference. Energy picked up from the radar antenna by the stays was causing an arc at the contact, and this was reradiated at a lower frequency corresponding to the natural period of the stays.

In many commercial broadcast receivers the antenna is coupled to the mixer, thus permitting radiation in the antenna from the local oscillator. The use of such receivers might produce radiation that could be detected by the enemy. Therefore all Navy receivers approved for shipboard use must have the oscillator separated from the antenna circuit by sufficient preselection and shielding to reduce radiation from the antenna to less than 400 micromicrowatts. The exact method of measuring this radiation is set forth in specifications provided by the Bureau of Ships. Electric

razors and other sources of interference not only cause difficulties in radio reception but also interrupt radio silence and render a ship vulnerable to attack.

It is the responsibility of the electronics officer to make certain that no equipment on the ship is emitting a radio signal on which an enemy direction finder can be trained.

SPECTRUM ANALYZER

A spectrum analyzer is an electronic test equipment which provides a visual indication of the frequency components (spectra) of an amplitude-modulated radio wave. The radio wave may be modulated by keying, by voice, by radar pulses, and so forth. In every case the resulting waves include a carrier frequency and its associated upper and lower sideband components. The pattern on the screen of a cathode-ray tube is a graph of signal voltage versus frequency. Ordinates represent peak voltage and abscissas represent frequency.

A simplified block diagram of one form of spectrum analyzer is shown in figure 13–18. The input signal is con-

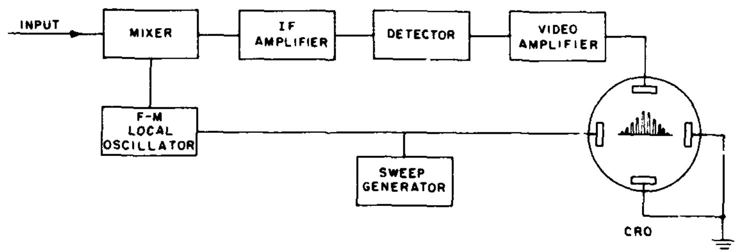


Figure 13–18.—Simplified block diagram of one form of spectrum analyzer.

verted in the mixer to the difference frequency (IF) between the input signal and the local oscillator. Any modulation present in the input signal is transferred to the intermediate frequency, removed from it at the detector, amplified by the video amplifier, and applied to the vertical deflection plates of the cathode-ray indicator tube. The action is similar to

that of a conventional superheterodyne receiver except that the local oscillator is frequency modulated. In other words, the frequency of the local oscillator is constantly being varied at a rate and to an extent that is determined by the sweep generator. The sweep generator has a saw-tooth waveform and sweeps the spot across the screen at a rate that is proportional to the rate of change of frequency of the local oscillator. Thus the horizontal position of the spot at any instant is proportional to the frequency of the applied signal at that instant and the vertical position of the spot is proportional to the amplitude of the signal.

Spectrum analyzers are frequently used to study the radio frequency spectrum produced when the carrier is amplitude modulated by a succession of rectangular pulses as in radar signals. A characteristic pattern of the pulse spectrum of a radar signal is shown in figure 13–19.

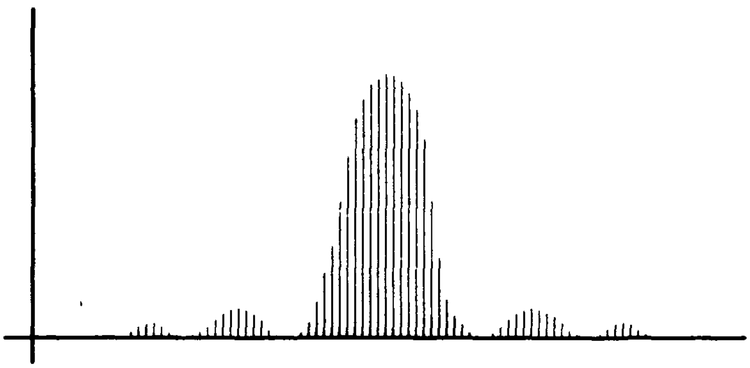


Figure 13–19.—Radar pulse spectrum.

The local oscillator of the spectrum analyzer superheterodyne receiver is swept in frequency at a rate that is proportional to the radar pulse recurrence frequency. The pattern is an envelope formed by the succession of pulses that are received during the time the spot is swept across the screen. In order to present sufficient detail in the screen pattern the sweep interval is made long enough to allow at least 50 pulses to occur for each sweep interval. Thus, the sweep speed

should not exceed one-fiftieth of the pulse-recurrence frequency. To avoid flicker, a long-persistence screen is used.

The TS–148/UP spectrum analyzer is a representative electronic test equipment used with aircraft radar and beacon equipment and provides a visual indication of the spectra of radio frequency oscillators within a range of 8,470 to 9,630 megacycles. A pictorial view of this spectrum analyzer control panel is shown in figure 13–20. This

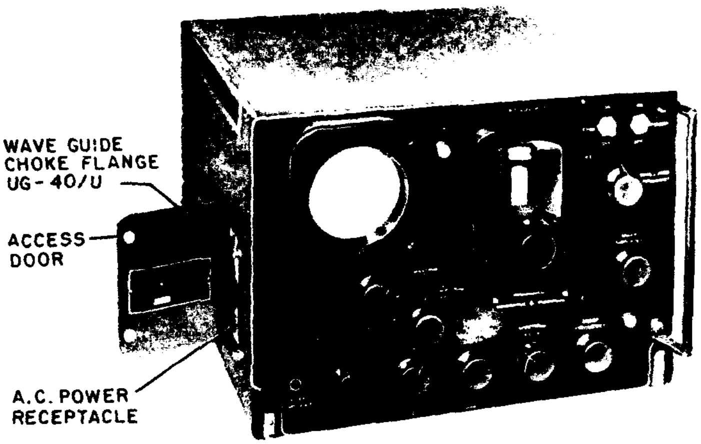


Figure 13–20.—Spectrum analyzer control panel, TS–148/UP.

analyzer is useful for observing spectra of pulsed magnetrons; measuring magnetron frequencies; tuning waveguides in a radar transmitter; checking frequency meters, TR boxes, and echo boxes; measuring the bandwidths of resonant cavities and pulse widths; and determining the distribution of useful transmitted power.

CAPACITANCE-INDUCTANCE-RESISTANCE BRIDGES

Capacitance, inductance, and resistance are measured for precision accuracy by means of alternating current bridges which are composed of capacitors, inductors, and resistors in a wide variety of combinations. These bridges operate on

the principle of the Wheatstone bridge in which an unknown resistance is balanced against known resistances. The unknown resistance is calculated in terms of the known resistances after the bridge has been balanced. One type of capacitance bridge circuit is shown in simplified form in figure 13–21. When the bridge is balanced by adjustment of

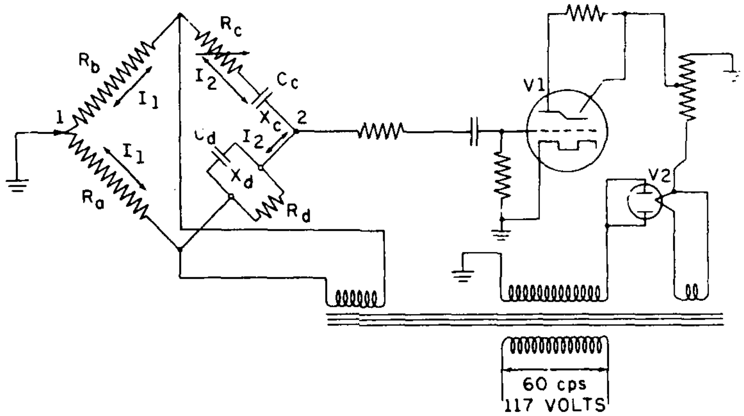


Figure 13–21.—Simplified schematic of capacity checker.

the two variable resistors, there is no a-c voltage developed across the input of the indicator tube, $V1$, and the shadow angle is maximum. Any slight unbalance produces an a-c voltage which in turn develops a grid-leak bias and lowers the plate current of $V1$ thus reducing the shadow angle.

The following relations exist when the bridge is balanced:

$$\frac{C_d}{C_e}=\frac{R_b}{R_a}=\frac{R_c}{R_d}, \qquad (13\text{–}1)$$

and

$$\omega^2=\frac{1}{R_dR_cC_dC_c}, \qquad (13\text{–}2)$$

where R_a, R_b, R_c, and R_d are the resistances indicated in the figure; C_e is the standard capacitance; and C_d the unknown capacitance. $\omega=2\pi f$, where f is the frequency of the voltage applied across the bridge.

In the basic Wheatstone bridge circuit using d-c voltages and simple resistances the balance is obtained when the voltage drops across the ratio arms are equal. In the a-c capacity bridge it is not sufficient to have equality of voltage drops in the ratio arms, but in addition the phase angle between current and voltage in the two arms containing the capacitors must be equal in order to obtain a balance. When a balance is obtained, the current in R_a is equal to that in R_b and the current in C_c is equal to the current in the parallel circuit of C_d and R_d.

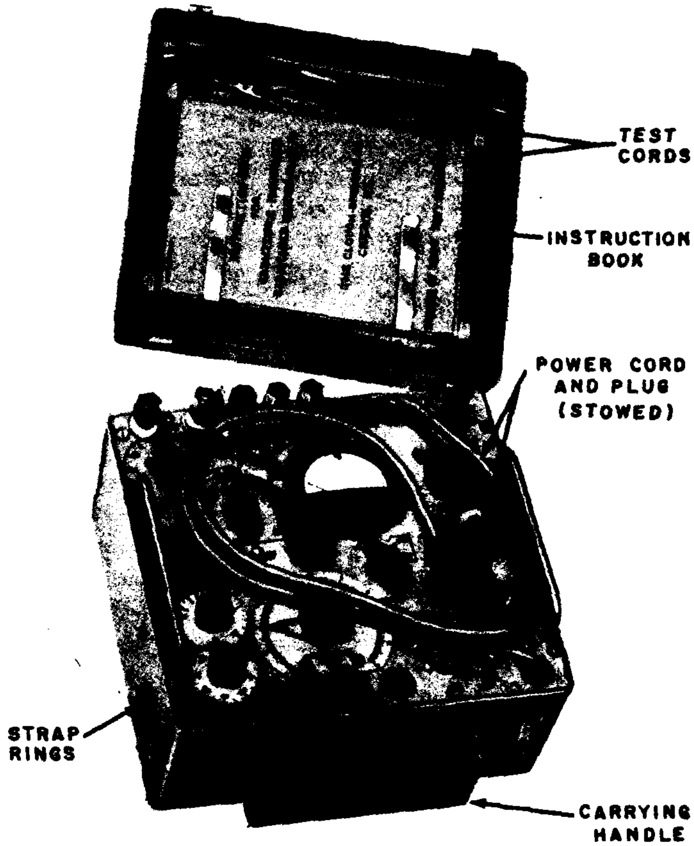


Figure 13-22.—Capacitance, inductance, resistance bridge, ZM-11/U.

The capacitance-inductance-resistance bridge, type ZM-11/U, shown in figure 13-22 is widely used to measure C, L, and R values in addition to special tests, such as the turns ratio of transformers and capacitor quality tests. This instrument is self-contained except for a source of line power and has its own source of 1,000-cps bridge current together with a sensitive bridge balance indicator, an adjustable source of direct current for electrolytic capacitor and insulation resistance testing, and a meter with suitable ranges for leakage current tests on electrolytic capacitors.

TUBE TESTERS

Two types of tube testers are in general use. One type, the EMISSION-TYPE TESTER, indicates the relative value of an electron tube in terms of its ability to emit electrons from the cathode. The second, and more accurate type, is the MUTUAL-CONDUCTANCE (or TRANSCONDUCTANCE) tube tester. This tube tester not only gives an indication of the electron emission, but also indicates the ability of the grid voltage to control the plate current.

Circuits for Tube Tests

Tube tester TV-3B/U shown in figure 13-23 is a portable tube tester of the dynamic mutual-conductance type designed to test and measure the mutual conductance of electron tubes of the receiving types and many of the smaller transmitting types.

A multimeter section, using the same indicator is also incorporated in the equipment to permit measurements of a-c and d-c volts, d-c milliamperes, and resistance and capacitance in a number of ranges.

Line voltage applied to the primary of the power-supply transformer is adjusted by means of a variable resistor in series with the primary power leads. The line adjustment switch connects the meter so that the meter deflection is proportional to the magnitude of the applied line voltage.

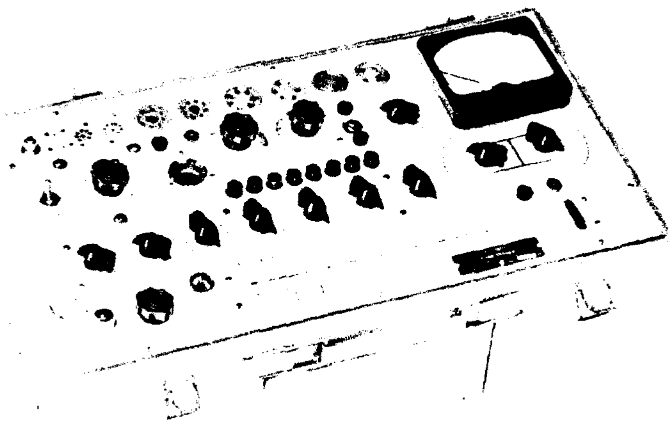


Figure 13-23.—Tube tester TV-3B/U.

The shorts switch connects the various tube elements to a voltage source in series with a neon lamp so that it glows if there is a short between the elements. The simplified circuit is shown in figure 13-24, A.

The noise-test jacks shown in figure 13-24, A, are connected to the antenna and ground posts of a radio receiver for the noise test. The short-test switch is turned through the various positions as the tube under test is tapped gently. Any intermittent disturbances between the electrodes cause momentary oscillations that are reproduced by the loudspeaker.

Rectifier and diode detector tubes are tested for emission as shown in the simplified circuit of figure 13-24, B. The tube being tested rectifies the alternating current and causes a pulsating direct current to flow through the meter. The current indicated by the meter is proportional to the electron emission of the tube. Rectifiers of the cold cathode type (such as the OZ4) require an a-c supply of 330 volts; whereas,

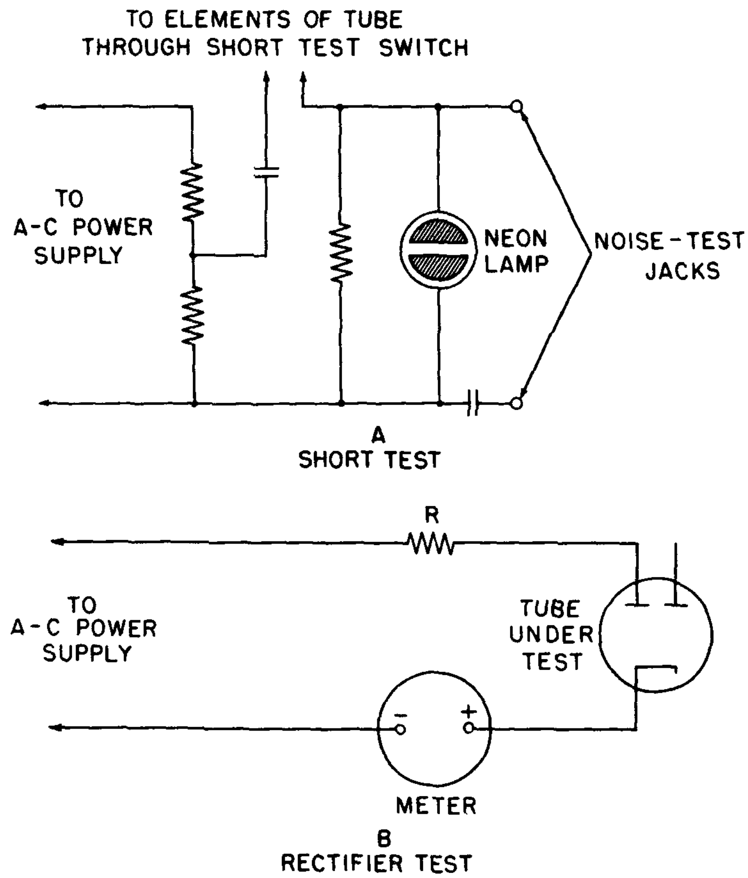


Figure 13–24.—Simplified short- and rectifier-test circuits.

diode detectors like the 6H6 require only about 20 volts. In each instance if there are two or more plates in the tube, each is tested separately.

The simplified circuits for the mutual-conductance test are shown in figure 13–25, A. The proper d-c grid voltage for the tube under test is supplied by $V2$. The a-c signal voltage is developed in $L5$ of the power transformer and acts in series with the grid bias.

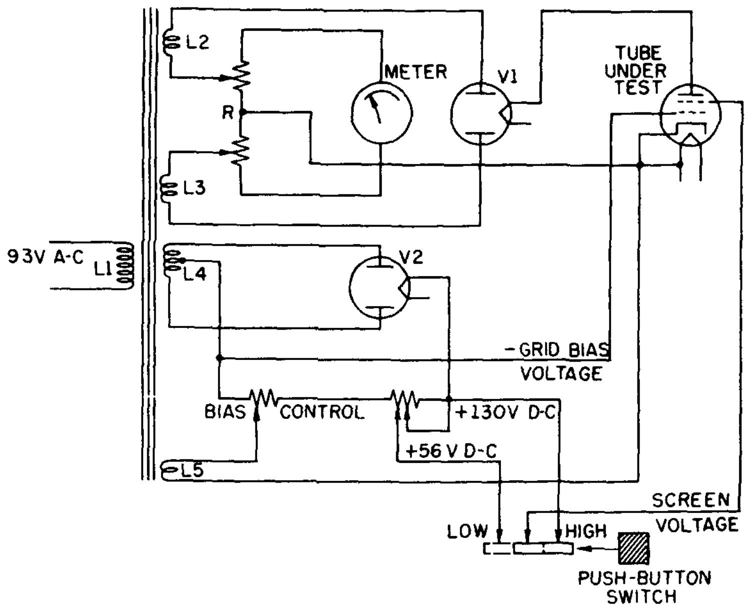


Figure 13–25.—Simplified mutual-conductance and gas-test circuits.

The plate voltage for the tube under test is supplied by *V1*. In the absence of a grid signal the a-c voltage on the d-c meter will cause no deflection because the currents in the two sections of *R* are equal and flow in opposite directions. However, when an a-c signal is applied to the grid of the tube under test, the plate current alternately increases and decreased through resistor *R* in phase with the grid signal voltage. Thus, the currents through the two sections of *R* become unbalanced and the voltage across the meter is equal to the difference in voltage across the two sections of *R*. The alternating deflecting force on the meter is thereby unbalanced and the indication is no longer zero. The meter indication is proportional to the average increase in plate current and is calibrated in micromhos.

The normal screen voltage of 130 volts is excessive for testing certain tubes. In such cases the screen may be

connected to a 56-volt source by means of the push-button switch as indicated in figure 13–25, A.

A simplified gas-test circuit is shown in figure 13–25, B. Depressing the gas-test push-button (pushed to the left) inserts resistor $R2$ in series with the grid. If the tube is gassy, reversed grid current will flow through $R2$. The drop across $R2$ is opposed to the grid bias voltage and plate current will increase.

Multimeter Section

The multimeter section of the tube tester includes a standard nonelectronic multirange volt-ohm-milliammeter. The theory of operation of these meters is included in training manuals for basic electricty.

VOLT-OHM-AMMETER—ELECTRONIC

The electronic volt-ohm-ammeter incorporates several measuring instruments within one enclosure. The ohmmeter and ammeter sections are similar to those described in manuals for basic electricity. The theory of operation of the ohmmeter and ammeter does not involve the electron tube and for that reason is not discussed in this training manual. The voltmeter section is described in this chapter because it involves the operation of the basic triode amplifier.

Voltmeter Errors

The ordinary voltmeter has several disadvantages that make it practically useless for measuring voltages in high-impedance circuits. For example, suppose that the plate voltage of a pentode amplifier is to be measured. When the meter is connected between the plate of the electron tube and ground, the meter current constitutes an appreciable part of the total current through the plate load resistor. Because of the shunting effect of the meter on the pentode, the plate voltage decreases as the current through the plate load resistor increases. As a result, an incorrect indication of plate voltage is obtained.

Before the voltmeter is connected, the plate current is limited by the effective resistance of the plate circuit and the plate voltage. If the tube has an effective resistance of 100,000 ohms, the plate load a resistance of 100,000 ohms, and the plate power supply is constant at 200 volts, then the plate current is $\frac{200}{200,000}$, or 0.001 ampere. The plate voltage is $0.001 \times 100,000$, or 100 volts.

Assume that the voltmeter used to measure the plate voltage of the tube has a sensitivity of 1,000 ohms per volt and that the range is from 0 to 250 volts. The meter will then have a resistance of 250,000 ohms. This resistance in parallel with the tube resistance of 100,000 ohms produces an effective resistance of 71,400 ohms in series with the plate load resistor. The total resistance across the B supply is therefore 171,400 ohms and the current through the plate load resistor is $\frac{200}{171,400}$, or 0.00117 ampere. Across the plate load resistor the voltage drop is $0.00117 \times 100,000$, or 117 volts and the plate-to-ground voltage on the tube is $200-117$, or 83 volts when the meter is connected, thus causing an error of 17 percent. The lower the sensitivity of the meter the greater this error will be.

A meter having a sensitivity of 20,000 ohms per volt and a 250-volt maximum scale reading would introduce an error of about 1 percent. However, in circuits where very high impedances are encountered, such as in grid circuits of electron tubes, even a meter of this sensitivity would impose too much of a load on the circuit.

Electron-Tube Voltmeter

Another limitation of the D'Arsonval a-c rectifier type voltmeter is the shunting effect at high frequencies of the relatively large meter rectifier capacitance. This shunting effect may be eliminated by replacing the usual metallic oxide rectifier with an electron-tube amplifier in which the plate circuit contains the d-c meter, and the voltage to be measured is applied to the grid circuit. Such a device is

called an ELECTRON-TUBE VOLTMETER. Voltages at frequencies up to and greater than 100 megacycles can be measured accurately with this type of meter.

THE INPUT IMPEDANCE IS LARGE, and therefore the current drawn from the circuit whose voltage is being measured is small and in most cases, negligible.

Simplified diagrams of the a-c and d-c electron-tube voltmeter sections of Navy model OBQ-4 volt-ohm-ammeter are shown in figure 13-26.

The operation of d-c amplifiers of the type used in electron-

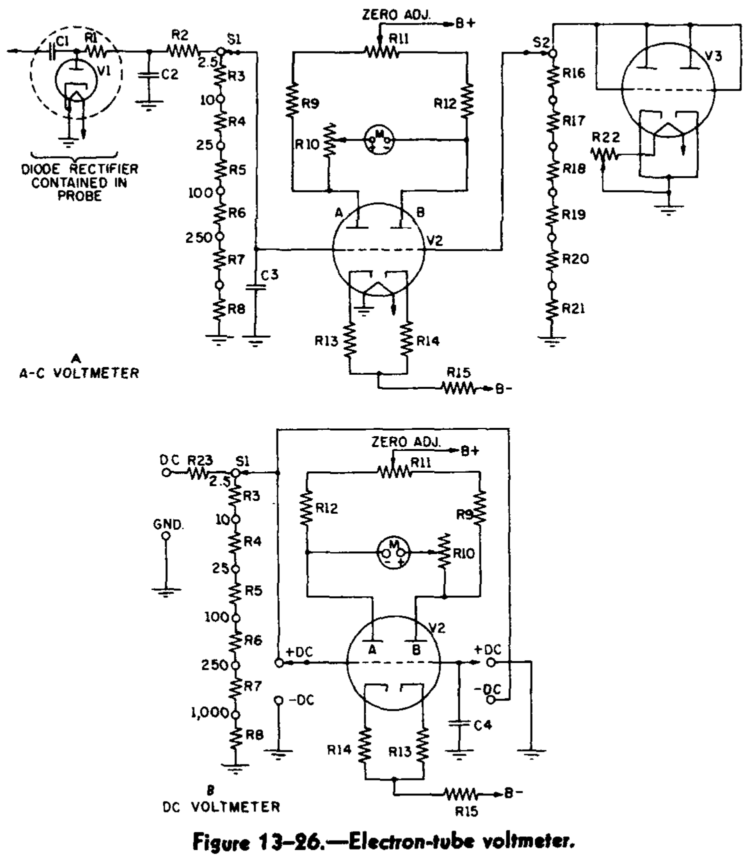


Figure 13-26.—Electron-tube voltmeter.

tube voltmeters is discussed in chapter 5 and may be reviewed for a better understanding of the operation of electron-tube voltmeters.

The a-c voltage to be measured is applied to the a-c probe (fig. 13–26, A). It is rectified by *V*1 and filtered by the *R-C* network in the probe.

The meter circuit is a balanced bridge network. When the input voltage between the probe and ground is zero, the bridge is balanced and the voltages across the two arms containing the plate load resistors of *V*2 are equal. Thus, the d-c meter indicates zero. If a voltage is applied between the probe and ground, the bridge becomes unbalanced and current flows through the meter. The meter is calibrated in rms volts. The input impedance is very high. At the lower frequencies the input capacitance is negligible, but as the frequency increases the input capacitance introduces an additional load on the circuit under test and causes an error in the meter reading.

The d-c electron-tube voltmeter circuit is shown in figure 13–26, B. The d-c voltage to be measured is applied between the d-c input terminal and ground. The d-c input voltage is therefore applied through *R*23 to the divider network feeding the grid of *V*2*A*. The grid of *V*2*B* is grounded. The meter is connected across a normally balanced bridge so that the application of the d-c voltage unbalances the bridge and causes the meter to deflect. The calibration is in d-c volts. Bias is obtained for *V*2*A* and *B* through the voltage drop across *R*13, *R*14, and *R*15. The cathodes are positive with respect to ground by an amount equal to the bias. Thus the grids are correspondingly negative with respect to the cathodes.

In figure 13–26, A, diode *V*1 causes a contact potential to be established across the voltage divider network connected to the grid of *V*2*A*. This voltage would unbalance the bridge. Therefore a similar contact potential is introduced across the grid of *V*2*B* from *V*3 and its associated voltage divider to balance the bridge before the a-c voltage to be measured is applied to the diode probe.

In figure 13–26, B, no diode probe is used, hence no contact potential is established so that $V3$ and its associated voltage divider network are omitted from the circuit.

TEST-TOOL SET

A test-tool set is provided electronics technicians for use as a test and repair set for general service work and emergency repair on electronic and electrical equipment. Representative of this type of equipment is test-tool set AN/USM–3, which includes a variety of tools and test equipment, as identified in figure 13–27. The set is supplied with the items fitted into a compact lightweight case in a manner such that all items are accessible and easily located.

The test-tool set includes a tube tester similar to the one previously described in this chapter. The set also includes a signal tracer which consists of a variable gain a-f amplifier and output meter. *Audio modulation on voltages* having frequencies of from 15 kc to 400 mc is detected by an accompanying test prod and amplified by the signal tracer. A telephone receiver, r-f cable assembly, and test prods are located in the accessory case to the left of the tube tester.

An interference generator is included to generate a-f and r-f voltages for test purposes. This unit is an aperiodic impulse buzzer-type generator housed in a probe case. Pressing the button at the top of the probe connects an enclosed battery with the buzzer. The buzzer frequency is approximately 2,000 cps with harmonics extending up to approximately 400 mc.

A voltage indicator probe is included in the test-tool kit. This unit consists of 2 meter elements so arranged that one indicates whether the line is a-c or d-c, and the second indicates the magnitude of the voltage. The second meter is calibrated in effective (rms) volts, which is about 10 percent higher than the average value when the voltage has a sine waveform.

An r-f probe is included which indicates the presence of

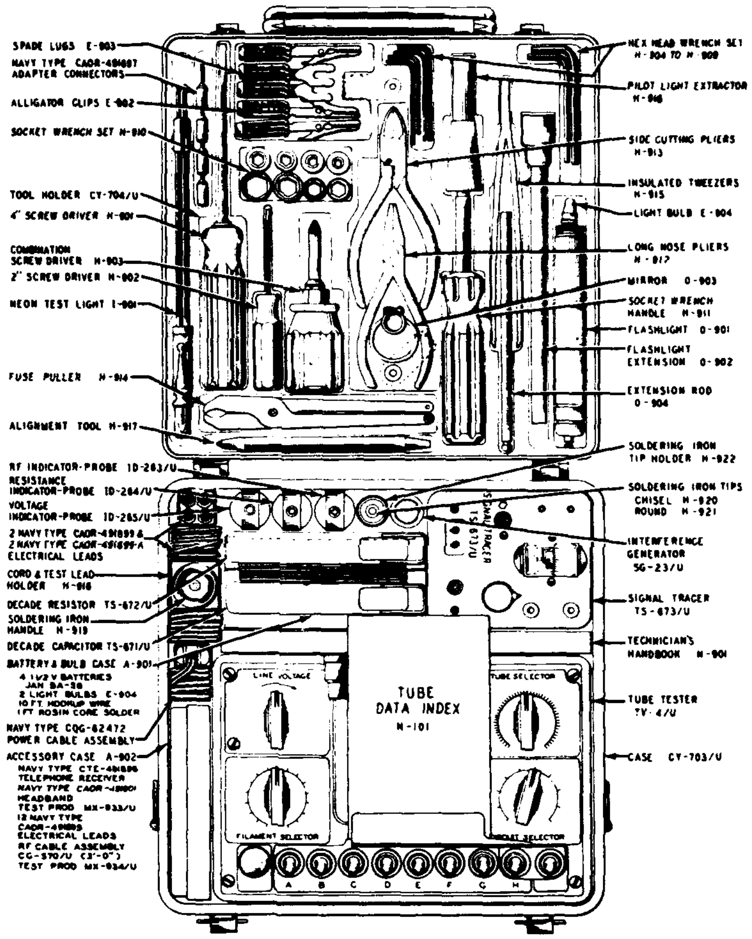


Figure 13-27.—Test-tool set AN/USM-3, identification of units.

r-f fields. The indicator contains two crystal rectifiers and microammeter with an isolating capacitor in the probe tip. This unit is designed to indicate the presence of intense r-f fields like those around transmitters and other r-f oscillators.

A resistance indicator probe is included in the test kit which indicates circuit continuity and resistance values

between 0 and 10,000 ohms. This probe is not a precision instrument and should be used to measure resistance only to a rough approximation.

A DECADE RESISTOR and a DECADE CAPACITOR are included in the test set. These units have a full range of values of resistance and capacitance required for test or temporary repair in trouble-shooting electronic equipment. The decade resistor can be used to obtain values of resistance from 1 ohm to 12 megohms. The decade capacitor can be used to obtain values of capacitance between 0.0001 and 48 microfarads.

Also included in the kit are an assortment of hand tools such as pliers, screwdrivers, wrenches, and tweezers; and a variety of accessories such as electrical leads, cables, soldering irons, mirror, extension rods, flashlight, and alignment tool.

QUIZ

1. What is the primary function of the cathode-ray oscilloscope as a test instrument?
2. How does the operator of a cathode-ray oscilloscope determine whether the pattern of a waveform is correct for the circuit under test?
3. What are the two types of deflection used in cathode-ray tubes?
4. In what type of deflection is the field that causes the deflection produced outside the cathode-ray tube?
5. In a cathode-ray tube, what controls the electron beam intensity?
6. What is the purpose of the second anode in a cathode-ray tube?
7. Why are the cathode-ray beam electrons not mutually repelled enough to defocus the beam?
8. How is focusing in an electrostatic-type cathode-ray tube generally controlled?
9. After the focus coil in an electromagnetic cathode-ray tube has been properly positioned, how is focusing accomplished?
10. What is meant by the electron beam deflection angle in a cathode-ray tube?
11. Define cathode-ray tube screen persistence.
12. How is a sine-wave voltage made to appear as a conventional sine curve on the cathode-ray screen?
13. Why are high-frequency signals sometimes applied directly to the vertical deflection plates of a cathode-ray oscilloscope rather than to the vertical amplifier input?
14. What is the purpose of the blanking pulse that is applied to the grid of the CRO through C9 in figure 13–6?
15. How may voltage waveforms be measured if they have magnitudes that are greater than those which the components within the oscilloscope can withstand?
16. Give two definitions of deflection sensitivity.
17. In a synchroscope, why is the sweep initiated before the signal pulse is applied to the vertical deflection plates?
18. State three uses of a synchroscope.
19. How may a multivibrator be used as an electronic switch?
20. Why are absorption wave meters not reliable for accurate measurements?
21. Why does grid current dip in a grid dip meter when the circuit under test has the same frequency as that of the meter?

22. Against what primary frequency standard are the reference crystals in frequency meters checked?
23. Why are secondary frequency standards or frequency meters not suitable for general service work as signal generators?
24. What are two applications of r-f signal generators?
25. What is the function of the r-f bridge in the LO-3 beat-frequency oscillator?
26. What are the essential components of a field-intensity meter?
27. What is the purpose of the shot-noise generator in the OF-2 interference-locating equipment?
28. In Navy receivers approved for shipboard use why must adequate preselection be used at the input?
29. What is the special function of the spectrum analyzer?
30. In an a-c capacity bridge, what conditions of voltage and phase angle must exist across the arms containing the capacitors to balance the bridge?
31. In addition to measuring L, C, and R values, what tests may be performed by the ZM-11/U bridge?
32. What is the essential difference between the emission-type tube tester and the transconductance-type tube tester?
33. What additional equipment is needed when the noise test is performed on a tube by means of the TV-3B/U tube tester?
34. For the mutual conductance test (fig. 13-25) what action unbalances the currents in the two sections of resistor R?
35. Why is the ordinary voltmeter useless for measuring the voltage in a high-impedance circuit?
36. Why is the usual a-c rectifier-type voltmeter inadequate for measuring high-frequency voltages?
37. What is the purpose of $V3$ in figure 13-26?

CHAPTER 14

INTRODUCTION TO RADAR

ELEMENTS OF RADAR

Definition

The word RADAR is formed from the initial letters of RAdio Detection And Ranging. It is an electronic device that may be used to detect the presence of objects like airplanes or ships in darkness, fog, or storm. In addition to indicating their presence, radar may be used to determine their bearing, distance, elevation, and speed; and to enable the operator to recognize their general character. It is one of the greatest scientific developments that has emerged from World War II. Its development, like the development of every other great invention, was mothered by necessity—that is, to detect the enemy before he detects us. The basic principles upon which its functioning depends are relatively simple, and the seemingly complicated series of electrical events encountered in radar can be resolved into a logical series of functions, which, taken individually, may be identified and understood.

Principles of Operation

SOUND-WAVE REFLECTION.—The principle upon which radar operates is very similar to the following principles of sound echoes or wave reflection. If a person shouts in the direction of a cliff, or some other sound-reflecting surface,

he hears his shout "return" from the direction of the cliff. What actually takes place is that the sound waves, generated by the shout, travel through the air until they strike the cliff. They are reflected or "bounced off," and some are returned to the originating spot, where the person is then able to hear the echo. Some time elapses between the instant the sound originates and the time when the echo is heard because sound waves travel through air at approximately 1,100 feet per second. The farther the person is from the cliff, the longer this time interval will be. If a person is 2,200 feet from the cliff when he shouts, 4 seconds elapse before he hears the echo—that is, 2 seconds for the sound waves to reach the cliff and 2 seconds for them to return.

If a directional device is built to transmit and receive sound, the principles of echo, together with a knowledge of the velocity of sound, can be used to determine the direction, distance, and height of the cliff shown in figure 14-1, A. A sound transmitter, which can generate pulses of sound energy, is so placed at the focus of the reflector that it radiates a beam of sound. The sound receiver is a highly directional microphone located inside a reflector (at its focal point, and facing the reflector) to increase the directional effect. The microphone is connected through an amplifier to a loudspeaker.

To determine the distance and direction, the transmitting and receiving apparatus is placed so that the line of travel of the transmitted sound beam and the received echo will very nearly coincide. They would coincide exactly if the same reflector could be used for both transmitting and receiving, as is done in radar systems. The apparatus (both the transmitter and receiver) is rotated until the maximum volume of echo is obtained. The horizontal distance to the cliff can then be computed by multiplying one-half of the elapsed time in seconds by the velocity of sound. This will be essentially the distance along the line *RA*. If the receiver has a circular scale that is marked off in degrees, and if it has been properly orientated with a

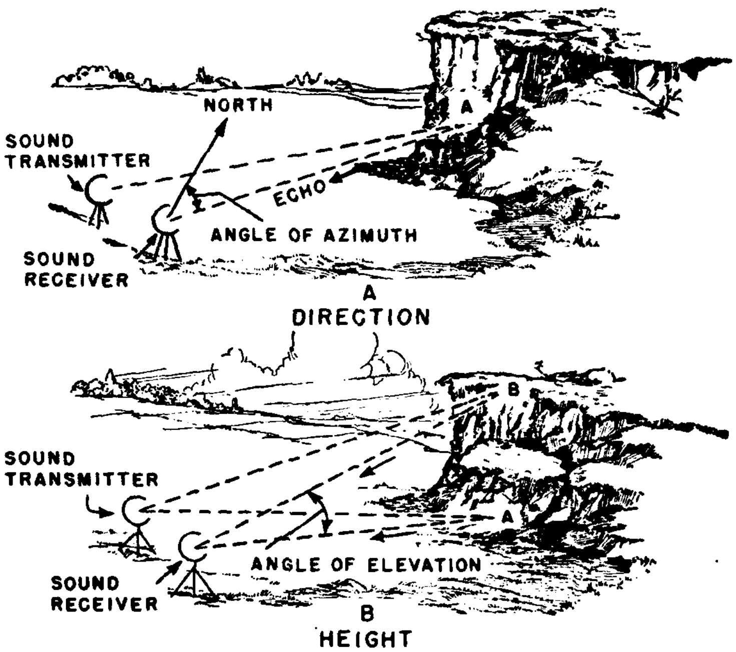


Figure 14-1.—Determination of direction and height.

compass, the direction or azimuth of the cliff can be found. Thus, if the angle indicated on the scale is 45°, the cliff is northeast from the receiver position.

To determine height (fig. 14-1, B), the transmitter and receiver antennas are tilted from the horizontal position (shown by dotted lines) while still pointing in the same direction. At first the echo is still heard, but the elapsed time is increased slightly. As the angle of elevation is increased, an angle is found where the echo disappears. This is the angle at which the sound is passing over the top of the cliff and is therefore not reflected back to the receiver. The angle at which the echo just disappears is such that the apparatus is pointing along solid line *RB*. If the receiver is equipped with a scale that permits a determination of the

angle of elevation, the height of the cliff, AB, can be calculated from this angle and either the distance RA or RB, by the use of one of the basic trigonometric ratios.

RADIO-WAVE REFLECTION.—All radar sets work on a principle very much like that described for sound waves. In radar sets, however, a radio wave of extremely high frequency is used instead of a sound wave (fig. 14-2). The

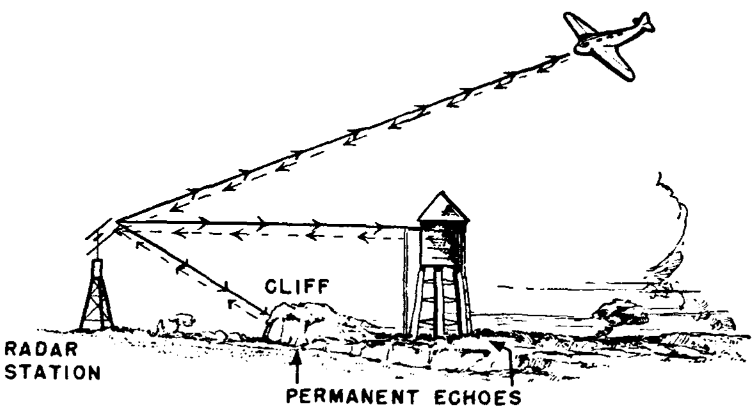


Figure 14-2.—Transmission and reflection of radar pulses.

energy sent out by a radar set is similar to that sent out by an ordinary radio transmitter.

The radar set, however, has one outstanding difference in that it picks up its own signals. It transmits a short pulse, and receives its echoes, then transmits another pulse and receives its echoes. This out-and-back cycle is repeated 60 to 4,000 times per second, depending on the design of the set. However, for Navy use the repetition rate is generally less than 1,000 cps. If the outgoing wave is sent into clear space, no energy is reflected back to the receiver. The wave and the energy that it carries simply travel out into space and are lost for all practical purposes.

If, however, the wave strikes an object such as an air-

plane, a ship, a building, or a hill, some of the energy is sent back as a reflected wave. If the object is a good conductor of electricity and is large compared to a quarter-wavelength of the transmitted energy, a strong echo (but only a very small fraction of the transmitted energy) is returned to the antenna. If the object is a poor conductor or is small, the reflected energy is small and the echo is weak.

Radio waves of extremely high frequencies travel in STRAIGHT LINES at the speed of light, or approximately 186,000 miles per second. Accordingly, there will be an extremely short time interval between the sending of the pulse and the reception of its echo. It is possible, however, to measure the interval of elapsed time between the transmitted and received pulse with great accuracy—even to one ten-millionth of a second (1×10^{-7} seconds). The forming, timing, and presentation of these pulses are accomplished by a number of special circuits and devices.

The directional antennas employed by radar equipment transmit and receive the energy in a fairly sharply defined beam. Therefore, when a signal is picked up, the antenna can be rotated until the received signal is maximum. The direction of the target is then determined by the position of the antenna.

The echoes received by the radar receiver appear as marks of light on an oscilloscope (often called "scope" for short). This scope may be marked with a scale of miles (or yards), or degrees, or both. Hence, from the position of a signal echo on the scope, an observer can tell the range and direction of the corresponding target.

Radar Methods

CONTINUOUS-WAVE METHOD.—The continuous-wave method of detecting a target makes use of the Doppler effect. The frequency of a radar echo is changed when the object which reflects the echo is moving toward or away from the radar transmitter. The change in frequency is known as the DOPPLER EFFECT. A similar effect at audible frequencies is recognized readily when the sound from the whistle of an

approaching train appears to the ear to increase in pitch. The opposite effect occurs when the train is moving away from the listener. The radar application of this effect permits a measurement of the difference in frequency between the transmitted and reflected energy and thus a determination of the presence and speed of the moving target. This method works well with fast-moving targets, but not well with those that are slow or stationary. C-w systems are therefore limited in present usage.

FREQUENCY-MODULATION METHOD.—In the frequency-modulation method the transmitted energy is varied continuously and periodically over a specified band of frequencies. The instantaneous frequency of the energy being radiated by the antenna therefore differs from the instantaneous frequency being received by the antenna, from the target. The frequency difference depends on the distance traveled and can be used as a measure of range. Moving targets produce a frequency shift in the returned signal because of the Doppler effect, and this affects the accuracy of range measurements.

PULSE-MODULATION METHOD.—In the pulse-modulation method the r-f energy is transmitted in short pulses the time duration of which may vary from 1 to 50 microseconds. If the transmitter is turned off before the reflected energy returns from the target the receiver can distinguish between the transmitted pulse and the reflected pulse. After all reflections have returned, the transmitter can be turned on again and the process repeated. The receiver output is applied to an indicator that measures the time interval between the transmission of the energy and its return as a reflection. Because the energy travels at a constant velocity, one-half the time interval becomes a measure of the distance traveled, or the range. Because this method does not depend on the relative frequency of the returned signal or on the motion of the target, difficulties experienced in the c-w and f-m methods are not present. The pulse-modulation method is used almost universally in military applications. Therefore, it will be the only method treated in this chapter.

Historical Development

One of the first observations of "radio echoes" was made in the United States in 1922 by Dr. A. H. Taylor at the Naval Research Laboratory. Dr. Taylor observed that a ship passing between a radio transmitter and receiver reflected some of the waves back toward the transmitter. Between 1922 and 1930 further tests proved the military value of this principle for the detection of objects that were hidden by smoke, fog, or darkness. During this same period Dr. Breit and Dr. Tuve of the Carnegie Institute published reports on the reflection of pulse transmission from electrified layers in the upper atmosphere. This led to the application of the principle to the detection of aircraft. Other countries carried on further experiments independently and with utmost secrecy. By 1936, the United States Army was engaged in the development of a radar warning system for coastal frontiers. By the end of 1940, mass production of radar equipment was under way. By September 1940, the British had developed radar to such a point that they were able to bring down great numbers of enemy airplanes, the guns being accurately controlled by radar systems. Beginning in 1941, British-American cooperation in the development of radar gave the United Nations the best radar equipment in the world. However, the enemy also made great strides in radar development. This was evidenced by the sinking of the British cruiser *Hood* by the German battleship *Bismarck*, by means of radar range finding, before the *Hood* could fire her second salvo.

Along with the development of radar went the development of effective countermeasures. Since 1941 great advances have been made in radar and in countermeasures in the various research and development centers throughout the country.

Uses of Radar

Radar equipments fall into three general classifications—(1) search, (2) fire control, and (3) identification.

Search radars are of two categories—(1) air search and

(2) surface search. These equipments are for general navigational use and early warning networks.

Fire control radars are confined more specifically to use with certain groups or types of gun batteries and are designed to fit the requirements of the battery with which they are used.

Identification radars are more commonly known as IFF ((identification, friend or foe), and are used to recognize ships and aircraft detected by the radar.

TYPES OF PRESENTATION.—To furnish usable intelligence, a radar set must have some type of visual presentation of the target echo for the operator to observe. Cathode-ray tubes are used for this purpose and are generally referred to as "scopes." Several types of data presentation have been developed to give the required information.

Type-A presentation is used to determine range, and the scope has short persistence. The echo causes a vertical displacement of the spot, the amplitude of which depends on the strength of the returned signal pulse. The point on the horizontal base line at which the vertical displacement occurs indicates the range. Type-A presentation is shown in figure 14–3, A.

Type-B presentation (fig. 14–3, B) indicates both range and azimuth angle. The vertical displacement of the echo signal indicates range, and the horizontal displacement of the echo signal indicates azimuth angle. This scope has long persistence.

Type-PPI (plan-position indication) presentation also indicates range and azimuth angle. In this type of presentation the electron beam is swept radially from the center of the scope to the outer edge as the deflection coils are moved around the neck of the tube in synchronization with the antenna rotation. In effect, a picture of surrounding objects is "painted" on the screen. This scope also has long persistence.

In type-A presentation the echo signal causes vertical deflection of the electron beam—in other words, it is DEFLEC-

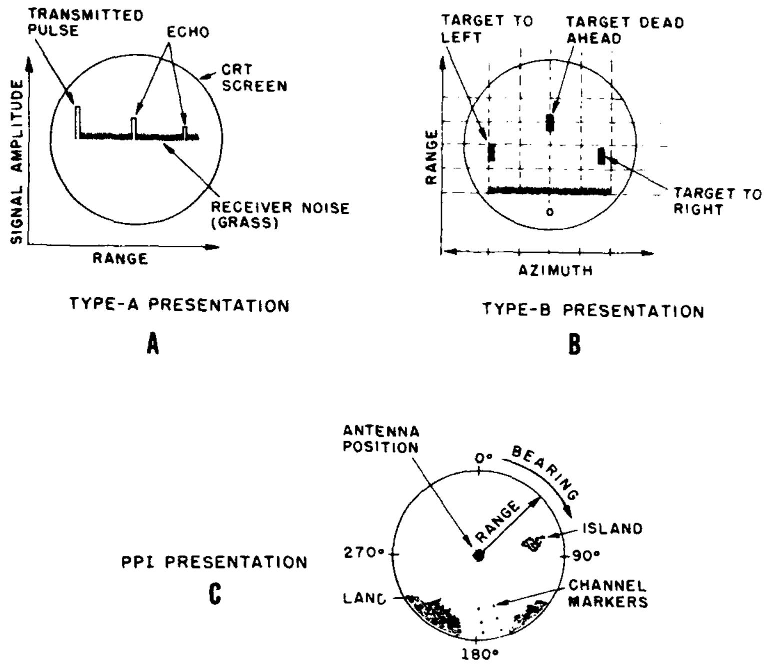


Figure 14-3.—Types of radar presentation.

TION MODULATED. In type-B and type-PPI presentation the echo signal makes the electron beam brighter. This is called INTENSITY MODULATION.

RANGE DETERMINATION.—The successful employment of pulse-modulated radar systems depends primarily on the ability to measure distance in terms of time and a knowledge of the velocity of light. Radio-frequency energy, once it has been radiated into space, continues to travel with a constant velocity. When it strikes a reflecting object there is no loss in time, but merely a redirecting of the energy. Its velocity is that of light, or, in terms of distance traveled per unit of time, 186,000 land miles per second, 162,000 nautical miles per second, or 328 yards per microsecond. This means that it takes approximately 6.1 microseconds

for radio energy to travel 1 nautical mile, or approximately 2,000 yards. All radar ranging is based on a flat figure of 2,000 yards per mile and, because the speed of light (and radio waves) is so great, microseconds (μs) are used for all time determinations.

This constant velocity of radio-frequency energy is applied in radar to determine range by measuring the time required for a pulse to travel to a target and return. The time lapse between the transmitted pulse and the echo return may be readily determined with the aid of the oscilloscope. For the purpose of illustrating how this may be done, assume that a target ship is 20 nautical miles away from the radar transmitter-receiver combination. Because radio energy travels 1 nautical mile in 6.1 microseconds, 122 microseconds will be required for the transmitted pulse to reach the target, or a total of 244 microseconds before the echo will return to the radar receiver.

The horizontal sweep frequency of the scope is adjusted so that it makes one complete sweep (from left to right) during the time the transmitted pulse is going to the target (maximum range) AND THE ECHO IS RETURNING TO THE RECEIVER. In other words, the time of one sweep is 244 microseconds, and the frequency is therefore approximately 4,098 cps. Assume that a translucent scale with uniform divisions in miles from 0 to 20 is placed over the face of the scope; and assume further that the extent of the sweep extends from the 0 mark to the 20-mile mark.

Figure 14–4 shows how the range is determined. In part ① the transmitted pulse is just leaving the antenna. A part of the generated energy is fed to the vertical deflection plates at the instant the pulse is transmitted and causes a vertical line (pip) to appear at the zero-mile mark on the scope.

In part ②, 61 microseconds later, the transmitted pulse has traveled 10 miles toward the target. The horizontal trace on the scope, however, has reached only the 5-mile mark—that is, one-half the distance the transmitted pulse

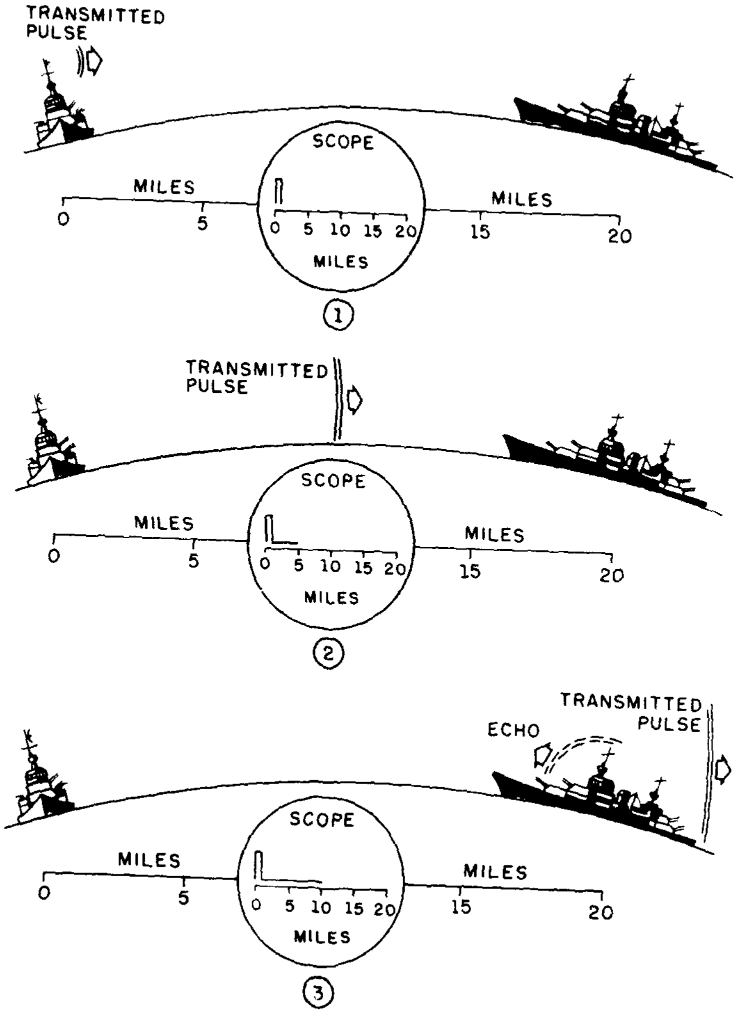


Figure 14-4.—Radar range determination (continued on page 664).

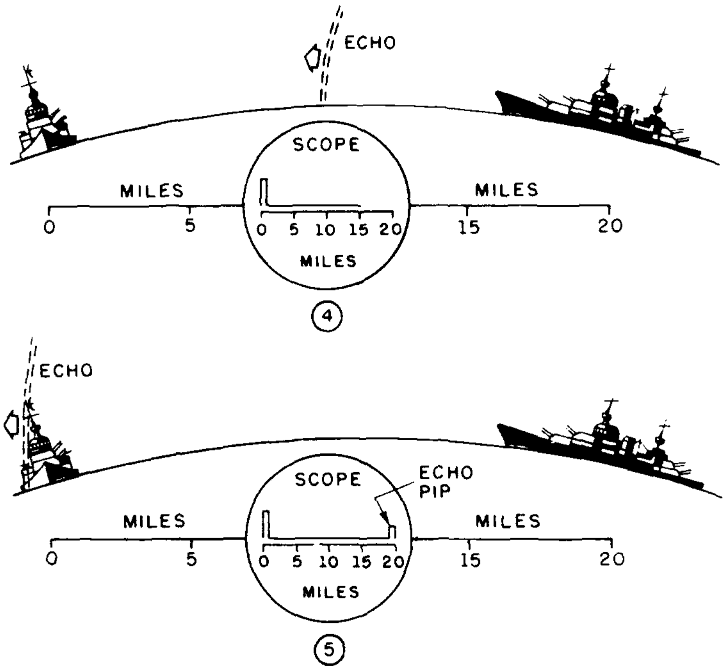


Figure 14-4.—Radar range determination—Continued.

has traveled (the sweep frequency is timed to indicate one-half the distance).

In part ③, 122 microseconds after the initial pulse left the transmitter, the transmitted pulse has reached the target, 20 miles away and the echo has started back. The scope reading is 10 miles.

In part ④, 183 microseconds after the initial pulse, the echo has returned half the distance from the target, and the scope reading is 15 miles.

In part ⑤, 244 microseconds after the initial pulse, the echo has returned to the receiving antenna. This relatively small amount of energy is amplified and applied to the vertical deflection plates, and an echo pip of smaller amplitude

than the initial pip is displayed on the scope at the 20-mile mark.

If two or more targets are in the path of the transmitted pulse each will return a portion of the incident energy as echoes. The target farthest away (assuming they are similar in size and type of material) will return the weakest echo.

In conjunction with the scope there is a handcrank and mechanical counter assembly which enables the operator to determine the range to a greater degree of accuracy. When a target is indicated on the base line the operator turns the handcrank to move the range indicator, or gate (fig. 14-5), to the target and then reads the range, in yards,

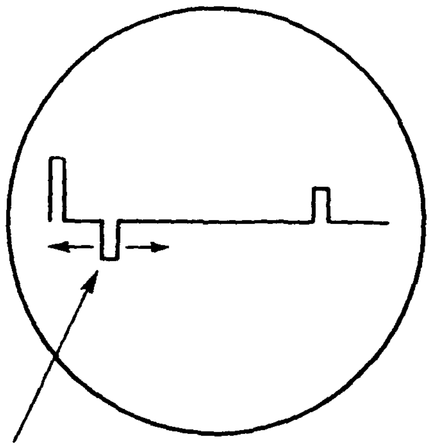


Figure 14-5.—Target gating.

directly from the counter assembly. This process is known as "gating the target."

Bearing determination.—The bearing (true or relative) of the target may be determined if the direction in which the directional antenna is pointing when the target is picked

up is known. Control and indicator systems have been devised that make this possible.

The measurement of the bearing of a target as "seen" by the radar is usually given as an angular position. The angle may be measured either from true north (called TRUE BEARING), or with respect to the heading of a vessel or aircraft containing the radar set (called RELATIVE BEARING). The angle at which the echo signal returns is measured by utilizing the directional characteristics of the radar antenna system. Radar antennas are constructed of radiating elements, reflectors, and directors to produce a single narrow beam of energy in one direction. The pattern produced in this manner permits the beaming of maximum energy in a desired direction. The transmitting pattern of an antenna system is also its receiving pattern. An antenna can therefore be used to transmit energy, to receive reflected energy, or to do both.

The simplest form of antenna for measuring azimuth or bearing is one that produces a single-lobe pattern. The system is mounted so that it can be rotated. Energy is directed across the region to be searched, by moving the beam back and forth in azimuth until a return signal is picked up. The position of the antenna is then adjusted to give maximum return signal.

Figure 14–6 shows the receiving pattern for a typical radar antenna. In this figure, relative signal strength is plotted against the angular position of the antenna with respect to the target. A maximum signal is received only when the axis of the lobe passes through the target. The sensitivity of this system depends on the angular width of the lobe pattern. The operator adjusts the position of the antenna system for maximum received signal. If the signal strength changes appreciably when the antenna is rotated through a small angle, the accuracy with which the on-target position can be selected is great. Thus, in figure 14–6, the relative signal strengths A and B have very little difference. If the energy is concentrated in a narrower beam, the difference is greater and the accuracy better.

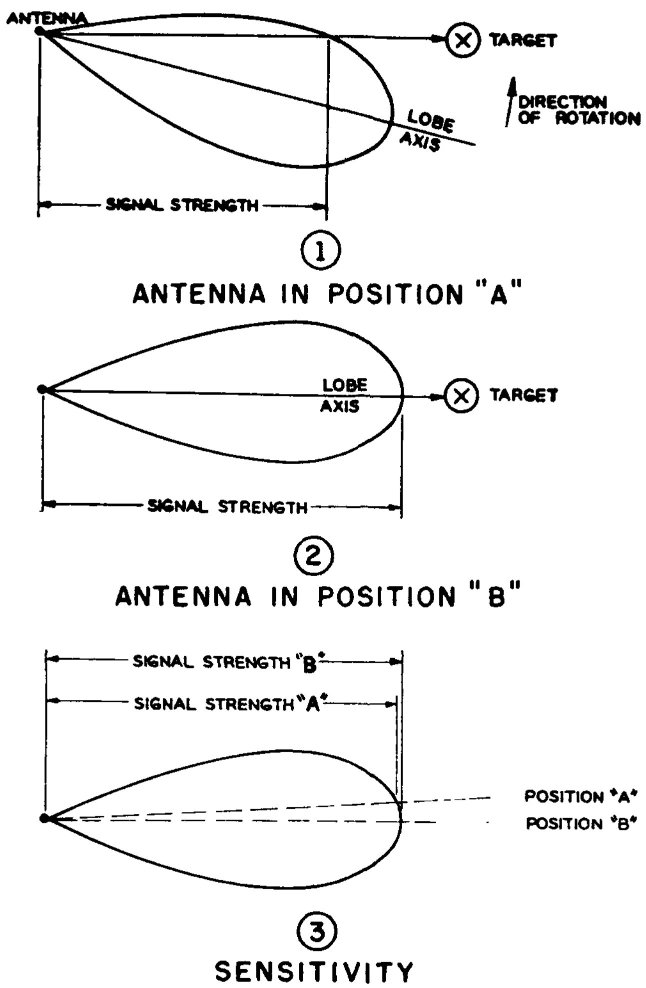
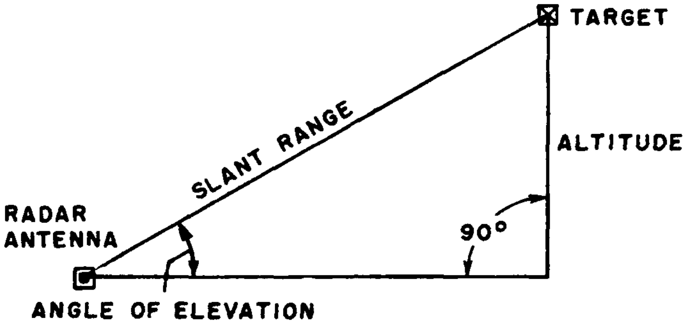


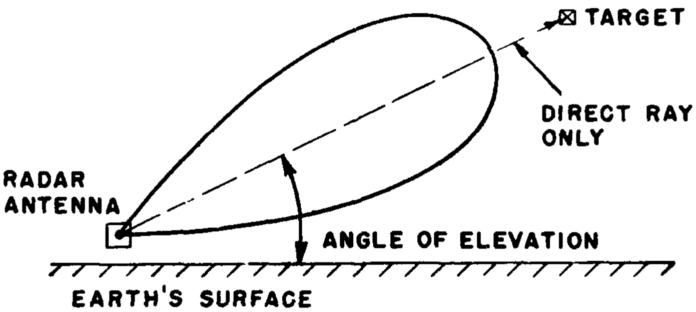
Figure 14-6.—Radar determination of azimuth or bearing.

ALTITUDE DETERMINATION.—The remaining dimension necessary to locate completely an object in space can be expressed either as an angle of elevation or as an altitude. If one is known, the other can be calculated from one of the

basic trigonometric ratios. A method of determining the angle of elevation or the altitude is shown in figure 14-7. The slant range (fig. 14-7, A) is obtained from the radar scope indication as the range to the target. The angle of elevation is that of the radar antenna (fig. 14-7, B). The altitude is equal to the slant range multiplied by the sine of the angle of elevation.



A
SLANT RANGE



B
ANGLE OF ELEVATION

Figure 14-7.—Radar determination of altitude.

In radar equipments with antennas that may be elevated, altitude determination by slant range is automatically computed electronically. In equipments (air search) where the antennas do not elevate, the altitude is automatically computed from the time lapse between the echoes that are returned directly to the receiver and those that return to the earth, which then reflects them back to the receiver.

PLAN POSITION INDICATOR.—The range scope has certain limitations when it is desired to know what is happening instantaneously in all directions because it indicates only the targets in the direction in which the antenna is instantaneously pointing.

A master PPI allows the radar operator to see the screen images of all objects surrounding his craft (within the range limitations of the equipments) because it displays a graphic plot of 360° of antenna rotation and has a screen of the necessary persistence to retain the targets visible after the antenna has rotated past the target bearing.

On most search radar equipments both range and PPI scopes are available to the operator.

The range scope presents the target information on a horizontal base line, as shown in figure 14–3, A. The PPI has a radial base line originating at the center of the screen (fig. 14–3, C) which indicates the physical antenna location, and this line follows the antenna rotation.

A view of a PPI scope is shown in figure 14–8. The bright spots on the screen are images of objects (ships, planes, land masses, etc.) in the vicinity of the craft carrying the PPI equipment. Around the outer edge of the scope are relative and true bearing circles. Spaced evenly across the face of the tube are range circles, calibrated in miles. Thus, from the position of the images, their approximate range and bearing may be determined from the scope. A particular object of interest may be singled out for more accurate ranging by referring to the range scope.

Another principal difference between the two systems (range and PPI) is the method of applying the signal to the scope. In the RANGE SCOPE the echo signal is amplified and

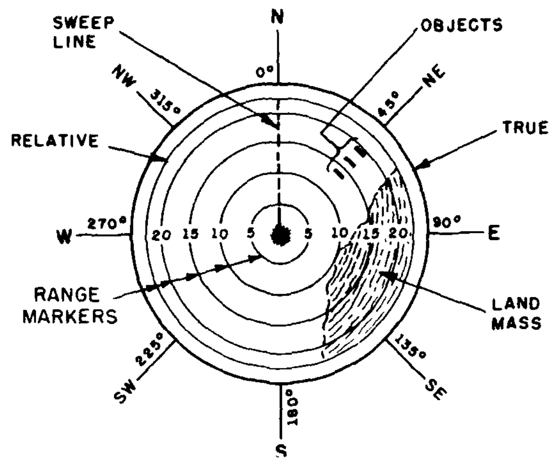


Figure 14-8.—PPI presentation.

applied to the vertical deflection plates in such a way as to produce a pip on the horizontal time-base line, on the screen. In the PPI SCOPE, the echo signal is amplified and applied to the control grid of the scope in such a way that the trace is brightened momentarily on the radial time-base line. If the intensity of the trace is kept sufficiently low, the scope will be essentially dark until an echo is received, and then the contrast will be very pronounced.

The PPI uses electromagnetic deflection instead of electrostatic deflection. Current flowing from the sweep generator through a single pair of electromagnets mounted across the neck of the tube at right angles to the axis of the tube causes the electron beam to be swept from the center of the tube to one edge and back again to the center.

The deflection electromagnets are mounted so that they can be rotated around the neck of the tube. The rotating assembly is synchronized with the antenna rotation so that when the antenna turns, the sweep trace is rotated about the screen at the same rate.

Thus, for example, in figure 14-8 when the antenna is

pointing in the NE direction the deflection magnets will force the beam across the screen from the center to the outer edge in the NE direction. The beam will be deflected across the screen many times during the course of a small angular rotation of the magnets. In this area on the screen the echoes from the three targets will cause three areas of intensification on the screen.

SEARCH RADAR.—Search radars used for general navigation and early warning nets do not require great precision in ranging or bearing, but do require the ability to locate targets at fairly long ranges. Therefore, they are normally designed with high power, wide beam angle, and fairly long pulse widths. Their target resolution (ability to accurately determine bearing and range) is approximately $\pm 2°$ and ± 200 yards. The 2° angle refers to the azimuth variation from the center of the target, and the 200 yards refers to the depth of the target.

FIRE CONTROL RADAR.—Fire control radars require only the necessary range for controlling the guns with which they are associated. However, they require precision target resolution. In this respect they are designed with low power, short pulses, and narrow beam angles. Their primary purpose is to furnish to the fire control systems accurate bearing and range of targets in order that the computers and other components of a gun-laying system may be fast, effective, and efficient.

IDENTIFICATION.—IFF (identification, friend or foe) equipment is a part of the radar in use today. This equipment, rather than being an actual radar, is an aid to radar. It has its own transmitter and receiver, and answers the all-important question of whether the target is enemy or friend. The IFF antenna and the radar antenna use a common reflector so that the two pulses are radiated from the same place.

AIRCRAFT RADAR.—Radars for aircraft operate on the same principle as that of shipboard radars except they are much smaller and lighter in weight. Both search and fire control radars are highly successful for aircraft use.

SPECIAL EQUIPMENT.—Special equipment, such as height-determining radar, may be used with radar to ascertain altitude. The special equipments are limited in their employment to one task.

FUNCTIONAL COMPONENTS

Radar systems now in existence vary greatly in detail. They may be very simple; or, if more accurate data are required, they may be highly refined. The principles of operation, however, are essentially the same for all systems. Thus a single basic radar system can be visualized in which the functional requirements hold equally well for all specific equipments.

In general, the degree of refinement of radar circuits increases with the frequency. The microwave region lends itself to a higher degree of precision in angular measurement, and for this reason modern radars operate at superhigh frequencies.

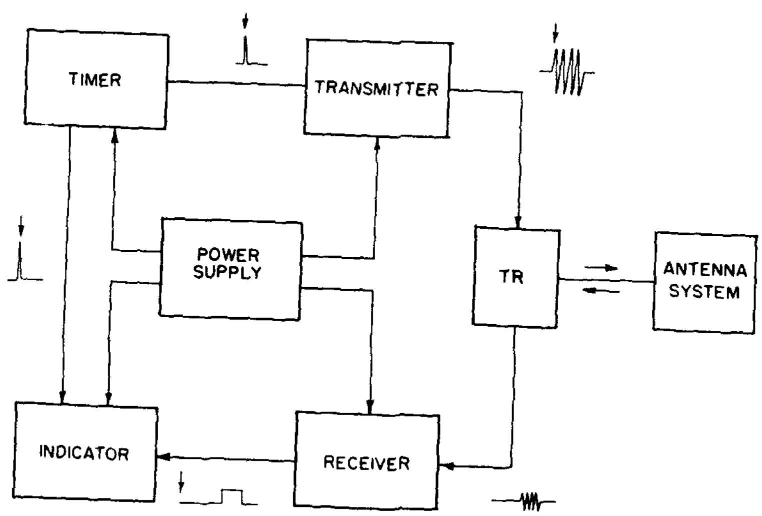


Figure 14-9.—Functional block diagram of a fundamental pulse-modulated radar system.

The functional breakdown of a pulse-modulated radar system generally includes six major components, as shown in the block diagram of figure 14–9. The components may be summarized as follows:

1. The TIMER (also called SYNCHRONIZER or KEYER) produces the synchronizing signals that trigger the transmitter the required number of times per second. It also triggers the indicator sweep and coordinates the other associated circuits.
2. The TRANSMITTER generates the r-f energy in the form of short, powerful pulses.
3. The ANTENNA SYSTEM takes the r-f energy from the transmitter, radiates it in a highly directional beam, receives any returning echoes, and passes these echoes to the receiver.
4. The RECEIVER amplifies the weak r-f pulses returned by the target and reproduces them as video pulses to be applied to the indicator.
5. The INDICATOR produces a visual indication of the echo pulses in a manner that furnishes the required information.
6. The POWER SUPPLY furnishes all a-c and d-c voltages necessary for the operation of the system components.

Before considering in more detail the action of the various functional components that make up a complete radar set, it is desirable to consider the radar system constants.

RADAR SYSTEM CONSTANTS

Any radar system has associated with it certain constants such as CARRIER FREQUENCY, PULSE-REPETITION FREQUENCY (the number of pulses sent out per second), PULSE WIDTH (in microseconds), and POWER RELATION (relationship of peak and average power). The choice of these constants for a particular system is determined by its tactical use, the accuracy required, the range to be covered, the practical physical size, and the problem of generating and receiving the signal.

Carrier Frequency

The carrier frequency is the frequency at which the r-f energy is generated. The principal factors influencing the selection of the carrier frequency are the desired directivity and the generation and reception of the necessary microwave r-f energy.

For the determination of direction and for the concentration of the transmitted energy so that a greater portion of it is useful, the antenna should be highly directive. The higher the carrier frequency, the shorter the wavelength and hence the smaller is the antenna array for a given sharpness of pattern, because the individual radiating element is normally a half-wave long. For an antenna array of a given physical size the pattern is sharper for a higher frequency.

The problem of generating and amplifying reasonable amounts of r-f energy at extremely high frequencies is complicated by the physical construction of the tubes to be used. The common triode becomes impractical and must be replaced by tubes of special design. Among these are such types as the "lighthouse" triode; grounded-grid triode; klystron; magnetron; and the "doorknob," "acorn," and "peanut" tubes.

In general, the modifications for extremely high-frequency operation are designed to reduce interelectrode capacitances, transit time, and stray inductance and capacitance in the tube leads. However, one of the problems involved in reducing the size of a tube is the reduction in the power rating of the tube. R-f generators, such as the magnetron, are designed to radiate a large amount of power during the relatively short pulse time.

Carrier frequencies may be of the order of 6,000 megacycles; however, the frequencies may extend down to 100 megacycles or up to 10,000 megacycles or above. Waveguides are of practical size only at the higher frequencies. At the receiver end, it is very difficult to amplify microwave signals; as a result, r-f amplifiers are not employed. Instead, the frequency of the incoming signal is mixed with that of a

local oscillator in a crystal mixer to produce a difference frequency called the INTERMEDIATE FREQUENCY (i-f). The intermediate frequency is low enough to be amplified in suitable i-f amplifier stages employing electron tubes.

Pulse-Repetition Frequency

Sufficient time must be allowed between each transmitted pulse for an echo to return from any target located within the maximum workable range of the system. Otherwise, the reception of the echoes from the more distant targets would be obscured by succeeding transmitted pulses. Therefore, the maximum range of a given equipment depends on the ratio of the resting time to the pulse width, provided the peak power is sufficient to return a usable echo. This necessary time interval fixes the highest pulse-repetition frequency that can be used.

When the antenna system is rotated at a constant speed, the beam of energy strikes a target for a relatively short time. During this time, a sufficient number of pulses of energy must be transmitted in order to return a signal that will produce the necessary indication on the oscilloscope screen. For example, an antenna rotated at 6 rpm having a pulse repetition frequency of 800 cps will produce approximately 22 pulses for each degree of antenna rotation. The persistence of the screen and the rotational speed of the antenna therefore determine the lowest pulse repetition frequency that can be used. In a system in which the indicator is operative during the entire interval between transmitted pulses, the repetition frequency must be very stable if accurate range measurement is desired. Because the oscilloscope screen will normally have a fairly long persistence, successive traces should appear in exactly the same position to avoid blurring.

Pulse Width

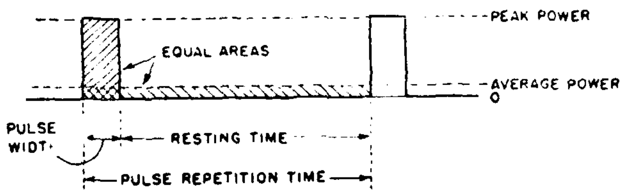
The minimum range at which a target can be detected is determined largely by the width of the transmitted pulse. If a target is so close to the transmitter that the echo is

returned to the receiver before the transmitter is turned off, the reception of the echo obviously will be masked by the transmitted pulse. For example, a pulse width of 1 μs will have a minimum range of 164 yards, meaning that a target within this range will not show, or will be "blocked out" on the screen. In this respect, equipments for "close in" ranging or navigation work use pulses of the order of 0.1 μs. For long-range equipment the pulse width is normally from 1 μs to 5 μs.

Power Relation

A radar transmitter generates r-f energy in the form of extremely short pulses and is turned off between pulses for comparatively long intervals. The useful power of the transmitter is that contained in the radiated pulses and is termed the PEAK POWER of the system. Power is normally measured as an average value over a relatively long period of time. Because the radar transmitter is resting for a time that is long with respect to the operating time, the average power delivered during one cycle of operation is relatively low compared with the peak power available during the pulse time.

A definite relationship exists between the average power dissipated over an extended period of time and the peak power developed during the pulse time. The over-all time



$$\text{DUTY CYCLE} = \frac{\text{AVERAGE POWER}}{\text{PEAK POWER}}$$

$$= \frac{\text{PULSE WIDTH}}{\text{PULSE-REPETITION TIME}}$$

Figure 14-10.—Relationship of peak and average power.

of one cycle of operation is the reciprocal of the pulse repetition frequency (PRF). Other factors remaining constant, the greater the pusle width, the higher will be the average power; and the longer the pulse-repetition time, the lower will be the average power. Thus,

$$\frac{\text{average power}}{\text{peak power}}=\frac{\text{pulse width}}{\text{pulse-repetition time}}.$$

These general relationships are shown in figure 14–10.

The operating cycle of the radar transmitter can be described in terms of the fraction of the total time that r-f energy is radiated. This time relationship is called the DUTY CYCLE and may be represented as

$$\text{duty cycle}=\frac{\text{pulse-width}}{\text{pulse-repetition time}}.$$

For example, the duty cycle of a radar having a pulse width of 2 microseconds and a pulse-repetition frequency of 500 cycles per second (pulse repetition time$=\frac{10^6}{500}$, or 2,000 microseconds is)

$$\text{duty cycle}=\frac{2}{2{,}000}=0.001.$$

Likewise, the ratio between the average power and peak power may be expressed in terms of the duty cycle. Thus,

$$\text{duty cycle}=\frac{\text{average power}}{\text{peak power}}.$$

In the foregoing example it may be assumed that the peak power is 200 kilowatts. Therefore, for a period of 2 microseconds a peak power of 200 kilowatts is supplied to the antenna, while for the remaining 1,998 microseconds the transmitter output is zero. Because

$$\text{average power}=\text{peak power}\times\text{duty cycle},$$

$$\text{average power}=200\times 0.001=0.2 \text{ kilowatts}.$$

High peak power is desirable in order to produce a strong echo over the maximum range of the equipment. Low average power enables the transmitter tubes and circuit components to be made smaller and more compact. Thus, it is advantageous to have a low duty cycle. The peak power that can be developed is dependent upon the interrelation between peak and average power, pulse width and pulse-repetition time, or duty cycle.

ELEMENTARY RADAR TRANSMITTER AND RECEIVER

Timer

The function of the timer is to ensure that all circuits connected with the radar system operate in a definite time relationship with each other and that the interval between pulses is of the proper length. In general, there are two practical methods of supplying the timing requirements—timing by means of a separate unit and timing within the transmitter.

A separate timing source may be used to give rigid control of the pulse-repetition frequency. In this case the source consists of any stable type of audio oscillator such as the Wein-bridge oscillator. The output is then applied to the necessary pulse-shaping circuits to produce the required timing pulse. Figure 14–11 shows in block form the functional components associated with the timer. These include the oscillator and other stages and components that are necessary to generate, shape, and amplify the waveform so that it may properly trigger the magnetron in the transmitter.

The oscillator generates a steady output at a given frequency (any frequency between 625 and 650 cps in the radar shown in the block diagram, and generally less than 1,000 cps), and this output establishes the PRF of the set. Frequency stability ensures that the range measurements will be accurate.

The sine-wave output of the oscillator is of the correct frequency but, it does not have the correct shape and its

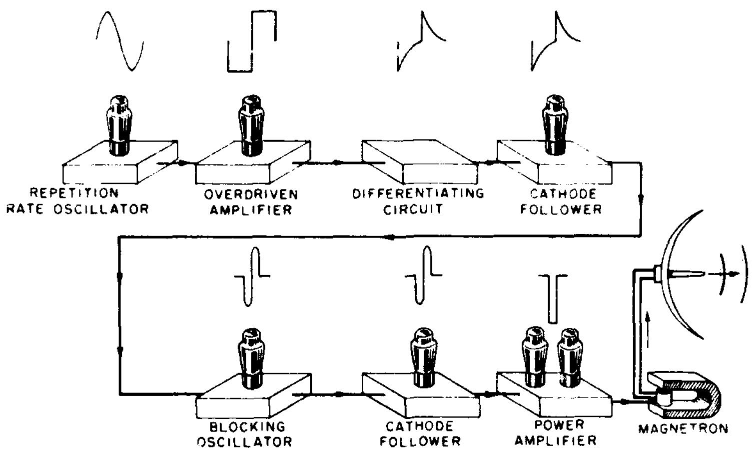


Figure 14–11.—Simplified block diagram of a modulator and transmitter.

amplitude is insufficient to fire the magnetron. Therefore, the signal is changed first into a square wave in the overdriven amplifier stage. The square wave is sharpened into a peaked wave in a differentiating circuit (a resistor and capacitor in series with the input, and the output taken across the resistor) and fed via a cathode follower to a blocking oscillator.

The blocking oscillator is triggered at the correct frequency by this peaked wave. The blocking oscillator generates the type of square wave needed by the magnetron, except that it is of insufficient amplitude.

The square-wave signal generated by the blocking oscillator is fed via a cathode follower to the power amplifier (preceded in actual circuits by driver amplifiers) where the square-wave pulse is amplified sufficiently to drive the magnetron. Only the negative portion of the pulse is used to drive the magnetron oscillator, and therefore the positive portion of the pulse is removed.

The magnetron goes into oscillation the moment it is triggered by the negative-going square wave from the power amplifier. The frequency of the magnetron oscillation may

be of the order of 6,500 megacycles. The width of the pulse is determined by the width of the negative-going pulse from the power amplifier and may be of the order of 1 microsecond. During the pulse, the power output may be of the order of 125 kw.

Transmitter

The transmitter is basically an r-f oscillator. It may be turned on and off by the negative-going pulse from the modulator. The radar oscillator (in this instance a magnetron) differs from the oscillators treated in chapter 7 in that it produces a much higher frequency and has a much higher power output. The higher frequency permits smaller waveguides and antennas to be used; and the higher power permits stronger echoes and a greater useful range.

Because of the superhigh frequencies in a radar set, buffers, frequency multipliers, and power output tubes following the magnetron would have little value in increasing the output power, and hence are not used in a radar set. The magnetron itself delivers 100 or more kilowatts of peak power to the transmission line, and yet it is relatively small. The more powerful sets are capable of putting out 1 megawatt (1,000 kw) of peak power. A simplified diagram of a magnetron is shown in figure 14–12, A. The magnetron is essentially a diode that has its plate at ground potential and its cathode at a high negative potential during the time it is oscillating. The diode is placed in a powerful magnetic field produced by a permanent magnet.

When a negative pulse is applied to the cathode, and there is no magnetic field present, electrons move in straight lines from the cathode to the plate, as shown in part ① of figure 14–12, B. When a weak magnetic field (part ②) is applied, the electron paths become curved; and as the magnetic field becomes stronger (parts ②, ③, ④, and ⑤), the electron paths become progressively more curved. Finally, the paths become so curved that the electrons are moving in closed circular orbits that miss the plate entirely, and no plate current flows. The plate is a copper block the internal

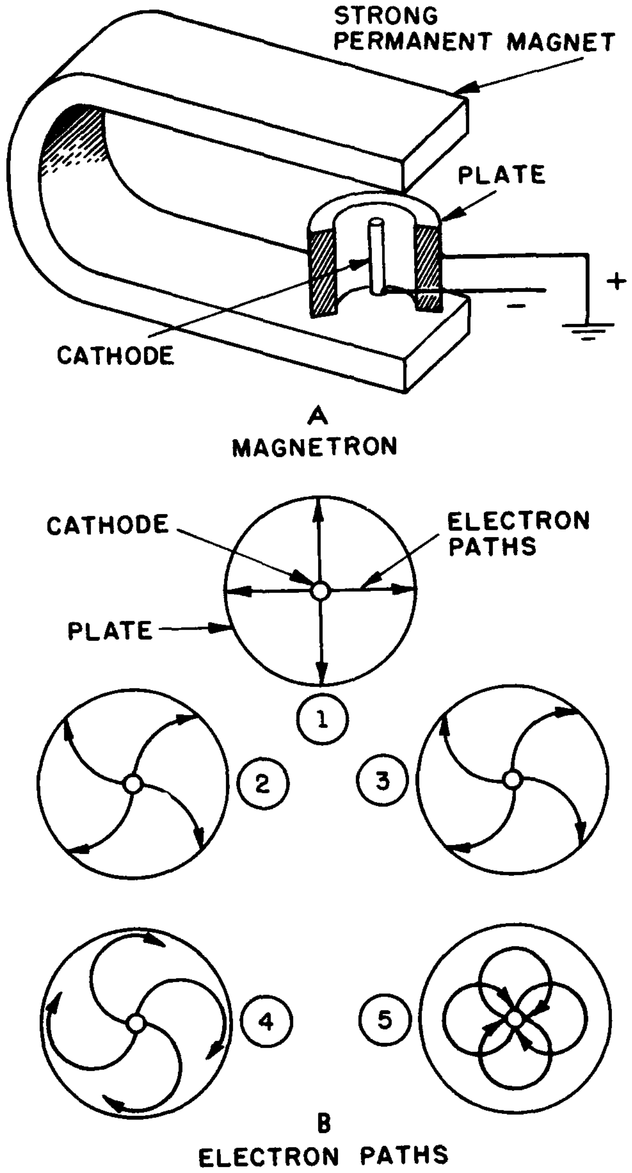


Figure 14–12.—Simplified diagram of a magnetron.

surface of which is separated into a number of segments by holes in the block that serve as tuned circuits. As the electrons move in circles past the plate segments they induce currents electrostatically in the walls of the holes. The energy of the magnetron output pulse is contained in the field associated with these currents. The smaller the circles the electrons make, the higher is the frequency of the oscillations. The frequency depends on the size of the cylinder, the strength of the magnetic field, and the difference in potential between the cathode and plate.

Energy is coupled out of the magnetron by means of a loop or probe; it is then transmitted to the antenna via a waveguide.

The tremendous peak power produced in short pulses by the magnetron requires high plate-to-cathode potential and high cathode emission. Because of the relatively long resting time between pulses, the problem of cooling is reduced and the physical size of the magnetron is not as large as would be expected from the peak power rating.

Transmitting and Receiving Antenna

The short powerful bursts of r-f energy produced by the magnetron are radiated in a narrow beam in the direction of the target by the action of the antenna. Likewise, echoes from the target are picked up by the antenna and passed on to the receiver and the display unit.

A general treatment of antennas is given in chapter 11; however, for the purposes of this chapter, only the reflector type of antenna system will be considered. Microwaves may be concentrated in essentially the desired pattern by means of reflectors in much the same way that light rays are brought into concentrated beams of light.

There are various types of reflectors for various purposes, and sometimes a combination of two or more reflectors is used to give greater versatility to the equipment. For example, one type of radar employs two types of reflectors, each having its own feed horn and each designed to serve a particular function. The assembly is shown in figure

14–13, A, and the patterns produced by the two antennas are shown in figure 14–13, B. The zenith reflector is de-

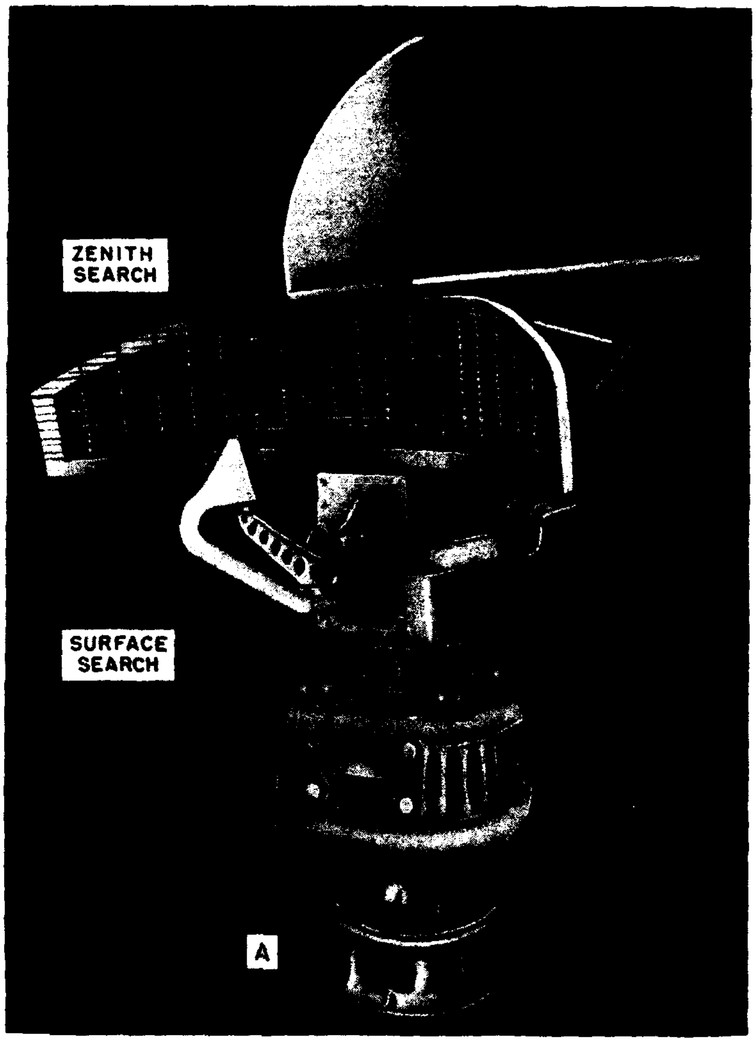


Figure 14–13.—Radar antenna assembly and antenna patterns **(continued on page 684).**

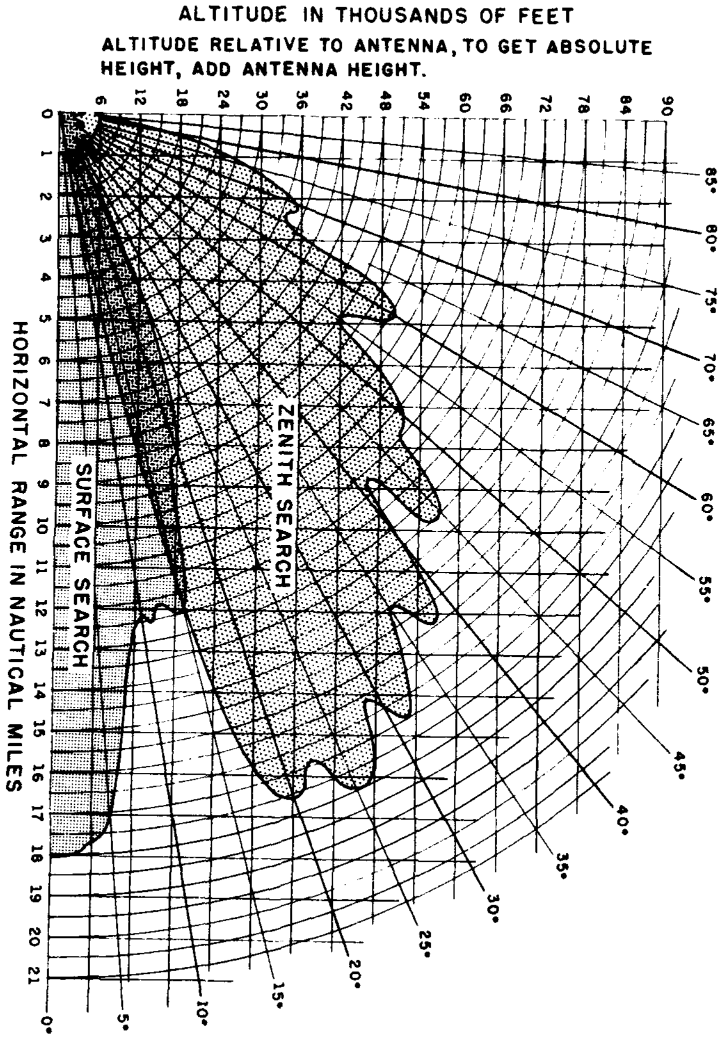


Figure 14-13.—Radar antenna assembly and antenna patterns—Continued.

signed to radiate the r-f energy at an angle above the surface of the earth where high-flying airplanes would probably be intercepted. The surface reflector radiates energy along the earth's surface, where ships or low-flying aircraft might be intercepted. The transmitting and receiving patterns for a given antenna are the same.

The waveguide is connected to the transmitter and receiver by means of the TR section. The function of this section is to close off electrically the waveguide to the receiver when the transmitter pulse is sent out and to close off electrically the waveguide to the transmitter when the echo pulse is received.

The TR section is necessary to prevent the strong pulse of outgoing energy from entering the receiver or the relatively weak returning echo from entering the transmitter. Otherwise the output pulse would damage the receiver and the echo would be lost in the transmitter.

Receiver

The radar receiver is essentially a special type of superheterodyne receiver. Its function is to receive the weak echoes from the antenna system, combine them in a crystal mixer (half-wave crystal rectifier) with the r-f signals from a local oscillator, amplify the resultant i-f signal, detect the pulse envelope, amplify the resulting d-c pulses, and feed them to the indicator. At the higher frequencies used in radar, it is not possible to use a stage of r-f amplification ahead of the mixer, and therefore the r-f signals are fed directly to the mixer.

In order to keep radar receivers in tune with their companion transmitters, a system of automatic frequency control is used in the receivers. Briefly, the system functions as follows: A small fraction of the r-f energy from the transmitter line is applied to a special automatic-frequency-control (a-f-c) mixer along with a small fraction of the r-f energy from the receiver local oscillator. The i-f energy resulting from the mixing of these two frequencies is amplified, rectified, and applied via control circuits in such a way

as to tune the local oscillator. If the i-f is of the correct frequency, the resulting direct voltage maintains the local oscillator at the correct oscillator frequency. If the i-f is too low in frequency the direct voltage applied to the local oscillator causes it to shift in frequency so that the i-f will be increased. If the i-f is too high, the oscillator frequency is shifted in the opposite direction.

The stability of operation is maintained in the microwave range of frequencies by careful design; and the over-all sensitivity of the receiver is greatly increased by the use of many i-f stages. Special types of tubes having low interelectrode capacitances also have been developed for use in local-oscillator and i-f stages. A block diagram of a radar receiver is shown in figure 14–14. As in communications receivers, the i-f signals in a radar receiver are fed to the

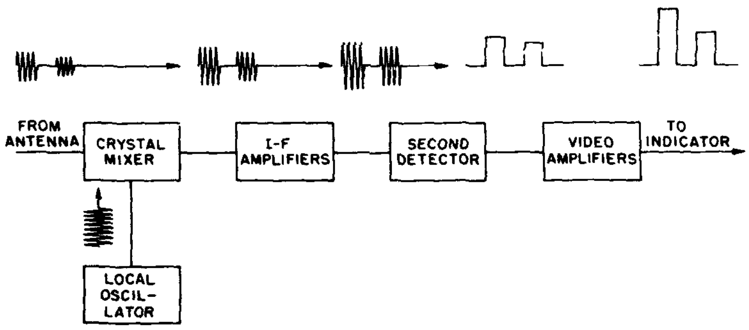


Figure 14–14.—Block diagram of a radar receiver.

second detector where the signal is rectified and the i-f component is removed. The remaining modulation pattern, consisting of d-c pulses, is fed to a video amplifier. In one type of presentation the output of the video amplifier is fed to the vertical deflection plates of an electrostatic-type cathode-ray tube. The amplitude of the vertical trace formed on the screen is proportional to the strength of the received signals. Simultaneously, a saw-tooth voltage is applied to the horizontal deflection plates in synchronism

with the transmitted pulse. The saw-tooth voltage provides a horizontal displacement that is proportional to range.

Radar video amplifiers have a frequency response similar to that of television video amplifiers. For example, a 5,000-mc search radar having a pulse duration of 1 microsecond would have a receiver bandwidth of 2 megacycles, and a fire control radar having the same pulse duration might have a receiver bandwidth of 5 megacycles.

Power Supply

In the functional diagram of the radar system (fig. 14–9) the power supply is represented as a single block. Functionally, this block is representative; however, it is unlikely that any one power supply could meet all the power requirements of a radar set. The distribution of the physical components of the system may be such as to make it impractical to lump the power-supply circuits into a single physical unit. Thus, different supplies are needed to meet the varying requirements of the system and must be designed accordingly. The power-supply function is performed, therefore, by various types of supplies distributed among the circuit components of the radar equipment. Power supplies are treated in chapter 3.

RADAR SPECIAL CIRCUITS

In order to GENERATE and RESHAPE the waves and to CONNECT one circuit to another in such a way as to cause the least disturbance to either circuit, several special circuits are needed. These circuits are listed and briefly described here, but are treated in detail in other chapters of this text and as needed in the rating books.

Generating Circuits

A generating circuit is one that produces oscillations of a given form and frequency.

REPETITION-RATE OSCILLATOR.—The repetition-rate oscillator, or timer, is a stable oscillator, capable of producing oscillations at an audio rate.

SAW-TOOTH GENERATOR.—The saw-tooth generator produces waves that resemble the shape of saw teeth. This type of generator may be operated continuously or it may be triggered by a sharp voltage pulse from another circuit. Saw-tooth waveforms are used to produce the sweep voltage on the electrostatic type of cathode-ray tube.

MULTIVIBRATOR.—The multivibrator is used to produce square waves of the desired frequency. Like the saw-tooth generator it may operate continuously, or it may be triggered into operation periodically by sharp voltage pulses. The square waves that trigger the magnetron are produced by a multivibrator that is in turn triggered by the timing pulse.

MAGNETRON.—The magnetron oscillator is also a generating circuit. It produces the high-frequency oscillations at sufficient power (during pulses) to properly "illuminate" the range area covered by the equipment.

LOCAL OSCILLATOR.—The local oscillator in the radar receiver generates high-frequency oscillations which, when mixed with the incoming pulse frequency, produce the i-f frequency. This type of oscillator at superhigh frequencies (3,000 mc and above) consists of a reflex velocity-modulated tube (klystron) tuned by cavity resonators. These are described in appropriate rating texts.

Reshaping Circuits

A reshaping circuit is one that takes the waveform from a generating circuit and shapes it according to the needs of the system.

LIMITING CIRCUIT.—Limiting, or clipping, circuits are employed to change the shape of the wave by clipping the top or bottom of the wave, or both. This may be accomplished by operating the grid with a small bias or a large bias or by overdriving a conventional amplifier. This type of amplifier is treated in chapter 12. The effects of too little or too much bias are treated in chapter 4.

CLAMPING CIRCUIT.—Clamping circuits, also called D-C RESTORERS, and BASELINE STABILIZERS, hold either amplitude extreme of a waveform to a given reference level of potential.

The diode clamper is used in radar to force each succeeding sweep to start at exactly the same spot on the radar scope. This eliminates "jitter" or "wobble" in the display pattern.

R-C DIFFERENTIATOR.—The *R-C* differentiator produces an output voltage the amplitude of which is proportional to the rate of change of the input voltage. Because a square wave has a fast rise and fall in voltage it may be applied to the input of an *R-C* differentiator to produce sharp voltage spikes across the output. These spikes may be used to trigger a multivibrator.

PHASE INVERTER.—Phase inverters (phase splitters) are used in oscilloscopes to provide from a single source two voltages that are 180° out of phase in order to provide a push-pull output from the horizontal amplifier stage. This type of circuit, together with some of its other uses, is treated in chapter 5.

Connecting Circuits

Connecting circuits are used to connect one circuit to another in such a way that minimum interference between the circuits will result. They are also used to enable a maximum transfer of energy or to accomplish some other desired result.

CATHODE FOLLOWER.—The cathode follower is a degenerative electron-tube circuit in which the inverse feedback is obtained by way of an unbypassed cathode resistor across which the output is taken. It is used to prevent interference between two circuits and as such becomes a "buffer" stage. These circuits are widely used as impedance-matching devices. Cathode followers are treated in chapter 5.

ELECTRONIC SWITCH.—The electronic switch is used to close, open, or change the operation of an electronic circuit. The electronic switch is very sensitive and is fast in operation. Thus, it can alternately connect one circuit to an oscilloscope, disconnect this circuit, and then connect a second circuit fast enough to present both waveforms simultaneously for a comparative study.

QUIZ

1. If a person shouts in the direction of a cliff and there is a 2-second interval before he hears the echo, how far is the cliff? (Assume the velocity of sound in air to be 1,100 ft/sec.)
2. What three measurements may be made with the apparatus shown in figure 14–1?
3. What effect is utilized in the continuous-wave radar method of detecting a target?
4. In the frequency-modulation radar method, upon what does the frequency difference of the outgoing and incoming signals depend?
5. Why are the problems experienced with the c-w and f-m radar methods not present in pulse radar?
6. What are the three general classifications of radar equipments?
7. What are the functions of the fighter-director radars?
8. What type of information is given in type-A presentation?
9. What type of information is given in type-B and PPI presentation?
10. How far (in yards) do radio waves travel in 1 microsecond?
11. How long (in microseconds) does it take a radio wave to travel 1 nautical mile (2,000 yards)?
12. What is the relation between the time for one sweep on the radar screen and the time for the transmitted pulse to travel to the target (maximum range) and return to the receiver?
13. Describe the process known as gating the target (fig. 14–5).
14. What is the effect on the accuracy of a radar of making the antenna beam angle narrower?
15. How may the altitude of a target be determined if the slant range and the angle of elevation are known?
16. What type of presentation has a radial time-base line?
17. How is the radar echo applied to the (1) range scope and (2) PPI scope?
18. In order to locate targets at long range, search radars have what special design features?
19. In order to obtain precision target resolution at short range fire control radars have what special design features?
20. What are the functions of the timer, or keyer, in a radar system?
21. What is the function of the transmitter in a radar system?
22. What is the relation between carrier frequency and the size of the radar antenna array for a given sharpness of pattern?

23. What is one of the problems involved when the size of an electron tube is reduced in order to reduce interelectrode capacitances and transit time?
24. Assuming sufficient peak power is available, upon what does the maximum range of a given radar equipment depend?
25. What determines the lowest pulse-repetition frequency that can be used with a given radar antenna and indicator system?
26. What determines the minimum range at which a target can be detected?
27. What is the term used to designate the ratio of the pulse width to the pulse-repetition time?
28. What type of pulse is needed to trigger the magnetron?
29. Upon what does the frequency of a magnetron depend?
30. Why does a magnetron having high peak power run relatively cool?
31. What is the function of the TR section?
32. Why is an r-f amplifier not used in a radar receiver?
33. How is the over-all sensitivity of a radar receiver improved?
34. Name five generating circuits that are applicable to radar special circuits.
35. What is the function of a reshaping circuit?

APPENDIX I

ANSWERS TO QUIZZES

CHAPTER 1

TUNED CIRCUITS

1. Because X_L and X_C are equal and opposite in polarity and therefore cancel each other.
2. The circuit impedance is a minimum at resonance, thereby allowing maximum current flow.
3. At a resonance, when the effective series resistance is low.
4. Because the currents in the two branches are approximately 180° out of phase and combine to produce a small resultant.
5. When the ratio of reactance to the inherent resistance in each unit is high.
6. To select the desired frequencies and to reject the undesired frequencies.
7. A vector quantity has magnitude and direction; a scalar quantity has only magnitude.
8. 5 times.
9. 90° clockwise from the 0° position, or $-90°$.
10. 5 units long and 180°.
11. 5 units of inductive reactance.
12. (a) $7+j5$; (b) $5+j$.
13. (a) -5; (b) $0.923+1.615j$.
14. (a) $1.732+j$; (b) $10\angle+53°$.
15. When they are parallel.
16. (a) $18\angle-30°$; (b) $2\angle+60°$.
17. Because the reactive components cancel.
18. The source voltage and the circuit current.
19. The ratio of the energy stored to the energy lost during the time the magnetic field is being established, and the ratio of X_L to R.
20. The ratio of the energy stored to the energy dissipated within the capacitor during the time the electrostatic field is being established, and the ratio of X_C to R.

21. It increases the circuit Q.
22. It increases the voltage gain.
23. The lower the Q, the flatter is the curve.
24. Above the resonant frequency the circuit acts like an inductor.
25. As a filter.
26. Because the nonenergy current in the capacitor is equal to the nonenergy current in the coil source and supplies only the energy component of current.
27. The $\frac{L}{C}$ ratio. (R being constant.)
28. Because the circuit acts like a capacitor in series with a resistor.
29. The impedance is reduced, the line current is increased, and the circuit Q is lowered. The phase angle is unaffected unless Q is less than 10.
30. Because in some electron-tube circuits it acts like a storage tank.
31. The high impedance of the series-tuned circuit and the low impedance of the parallel-tuned circuit offered to the frequencies outside the band in the vicinity of resonance.
32. The parallel-tuned circuit offers a high impedance and the series-tuned circuit offers a low impedance to the band in the vicinity of resonance.
33. To eliminate (trap out) from a circuit a given frequency or band of frequencies.
34. It reduces the primary current.
35. When the coupling is small and the plate resistance is large with respect to the coupled impedance.
36. The coefficient of coupling and the Q of the circuits.

CHAPTER 2

OPERATING PRINCIPLES OF THE ELECTRON TUBE

1. To convert currents and voltages from one waveform to another, to amplify weak signals, and to generate high-frequency currents.
2. Heat.
3. The frequency of the incident radiant energy.
4. Their great durability.
5. Medium power tubes with plate voltage between 500 and 5,000 volts.

6. Oxide-coated emitter.
7. Too much power is required for heating purposes.
8. To prevent oxidation of the cathode and heating element and to permit the flow of current from cathode to plate without colliding with gas particles.
9. The plate voltage at which all of the electrons transmitted by the cathode are attracted to the plate.
10. By the voltage between plate and cathode.
11. Oxide-coated emitter.
12. As rectifiers.
13. The cutoff bias.
14. When grid current flows.
15. Dynamic characteristics are obtained with a-c components present, as in actual operation; static characteristics are obtained with d-c components only.
16. Mu (μ).
17. A-c plate resistance (r_p).
18. $\mu = r_p g_m$.
19. g_m.
20. Distortion is produced in the output.
21. By reducing the physical dimensions and spacing of the electrodes and often not terminating them in conventional tube bases.
22. In tetrodes neutralization is usually unnecessary, plate current is practically independent of plate-voltage variations, and the amplification factor and the transconductance are higher.
23. Negative resistance.
24. To suppress secondary emission.
25. The space charge between the screen grid and plate repels them.
26. Variable-mu tubes.
27. Interelectrode capacitance and electron transit time.
28. By decreasing the plate potential below the deionization potential.
29. Higher plate current and lower internal resistance when conducting.
30. The plate supply.
31. Arcback, positive-ion bombardment of the cathode, and lack of sufficient deionizing time.
32. To give the mercury sufficient time to vaporize and thus to prevent bombardment of the cathode with ions that have acquired an excessive charge.
33. When conduction starts, the thyratron grid loses control.
34. As a tuning or balance indicator.

CHAPTER 3

POWER SUPPLIES FOR ELECTRONIC EQUIPMENTS

1. The B-supply is always d-c and supplies high-voltage at low current; the A-supply may supply either a-c or d-c and furnishes low voltage at relatively high current.
2. To minimize hum caused by the a-c heater component modulating the space current.
3. The heating and emitting elements are electrically insulated and therefore the a-c heater component is not coupled to the signal circuits.
4. High-vacuum tube.
5. Maximum peak plate current and maximum inverse peak plate voltage rating.
6. Because of the lower voltage drop across the mercury-vapor tube resulting in less power loss in the tube.
7. 15 volts.
8. The oxide coating is thin and therefore may be easily punctured.
9. By connecting a number of units in series.
10. Electrons flow from negative to positive through *R*, making the cathode positive with respect to ground.
11. 141.4 volts and 127.2 volts respectively.
12. The d-c load current flows through the two halves of the secondary in opposite directions.
13. In the bridge rectifier the output voltage is nearly twice as high.
14. Three filament transformer windings that are insulated to withstand the full load voltage between any two of them are needed for the tubes.
15. Because the charge-path is through the conducting diode and low-resistance transformer secondary, and the discharge-path is through the relatively high resistance of the load, *R*.
16. The direct voltage falls too much; and the surge of current when the capacitor charges may damage the gas tube.
17. The impedance of the input choke prevents the capacitor from charging up to the peak a-c voltage.
18. In the first (input) component.
19. Because current flow is intermittant and the choke has no effect except when current flows.
20. The voltage drop within the power supply caused by the flow of load current through the internal resistance of the power supply.

21. To prevent the capacitor from charging to the peak value fo the a-c voltage and thereby causing poor regulation.
22. Increased.
23. It is made worse.
24. The impedance of the tube varies inversely with the degree of ionization and with the amount of current drawn through the tube, so that the voltage drop across the tube and the load remains constant.
25. It increases.
26. The decrease in bias and effective resistance of the series triode.
27. The high amplification of the pentode makes the regulator more sensitive to small variations in load voltage or in input voltage, hence a more stable load voltage is maintained.
28. To apply a minimum fixed load to the filter and to discharge the filter capacitors when the power supply is turned off.
29. To supply a variety of output voltages.
30. To supply both positive and negative potentials with respect to ground.
31. The output voltage across that part of the divider is decreased and the division of voltages is altered.
32. When *D2* is nonconducting, no current is drawn from the input circuit and the size of *C2* limits the load current.
33. To maintain a steady voltage across the resistor at the lowest operating frequency.
34. To protect the tubes and circuits should other systems of bias fail.
35. To produce a higher direct voltage for the plates and screens of electron tubes.
36. Low-voltage d-c is converted into high-voltage a-c or d-c.
37. The synchronous vibrator does not require a separate rectifier, the nonsynchronous type does.
38. To change direct current into alternating current.

CHAPTER 4

INTRODUCTION TO ELECTRON-TUBE AMPLIFIERS

1. The d-c component.
2. By having a step-up or a step-down turns ratio.
3. To increase the relatively low amplitude of an input signal on the grid to a relatively high amplitude in the output (plate) circuit.

4. Voltage gain$=\frac{\text{signal voltage output}}{\text{signal voltage input}}$.
5. The ratio of the output power to the input grid driving power.
6. The ratio of useful output power to d-c input power in the plate circuit.
7. The ratio of the output power in watts to the square of the effective value of grid signal voltage.
8. The entire cycle.
9. Each tube supplies that half of the waveform not supplied by the other, thus giving a true reproduction of the input signal.
10. It flows for more than half but less than the full cycle.
11. As r-f power amplifiers in transmitters.
12. To make the response flat over a wide range of frequencies and to keep time-delay distortion within a certain minimum value at the high- and low-frequency ends of the spectrum.
13. The feedback coupling between plate and grid is removed and the grid-plate interelectrode capacitance is placed effectively in parallel with the load.
14. It has less input capacitance.
15. The ratio of the bandwidth to the center frequency.
16. Frequency distortion.
17. Phase distortion.
18. Amplitude (nonlinear) distortion.
19. Tube noise.
20. Reduced amplification.
21. The increased capacitive reactance of the interstage coupling capacitor.
22. Shunting the load with C_o, C_i, and C_d which increases the internal voltage loss in r_p, and reduces e_{out}.
23. It is not affected because the reactance of the series coupling capacitor is low and that of the shunting capacitances is high.
24. (1) The voltage gain may exceed mu. (2) The grid of the output stage is isolated from the B-supply voltage without the need for a blocking capacitor. (3) The d-c component of voltage acting in the plate circuit is less than when *R-C* coupling is used.
25. The decrease in reactance of the transformer primary inductance.
26. Because of the decreased reactance of the shunting capacitance across which the output voltage is developed, and because of the increased reactance of the leakage inductance in series with the output.
27. Positive.

28. Uniform frequency response over a wide frequency range, and suitability for amplifying pulse signals.

CHAPTER 5

ELECTRON-TUBE AMPLIFIER CIRCUITS

1. (a) Meter, relay and counter; (b) gain control of an amplifier, and frequency control of an oscillator.
2. Both gain and selectivity are increased.
3. The undesirable distortion is increased.
4. The nonlinear distortion is reduced.
5. Because it must occur in the plate circuit of the stage across which feedback is to be applied in order to separate the distortion from the desired signal.
6. In the high-level stages.
7. The high frequencies.
8. The gain is reduced.
9. The cathode resistor bypass capacitor.
10. They are 180° out of phase.
11. They are both high.
12. In series.
13. Directly.
14. It increases the Q.
15. The response curve is sharply peaked, and the pass band is very narrow.
16. The response at resonance is reduced and humps occur at both sides of resonance.
17. It lowers the circuit Q.
18. To introduce degeneration to overcome frequency distortion.
19. To resonate with the interelectrode and distributed capacitances and thus extend the high-frequency limit.
20. To increase the effective load impedance at the low frequencies and thus counteract the normal drop at these frequencies.
21. The gain increases with the capacitance.
22. Less.
23. Because the output voltage follows (is in phase with) the input voltage.
24. It decreases both.
25. Low.

26. Such an omission increases the low-frequency response.
27. To produce from a single waveform two waveforms that have exactly opposite instantaneous polarities.
28. Distortions and losses inherent in transformers.
29. (1) Degenerative feedback and (2) voltage division.
30. To reduce the input of $V2$ to the same value as the input of $V1$ so that the output voltage of $V2$ will equal that of $V1$.

CHAPTER 6

AUDIO POWER AMPLIFIER

1. To deliver power to a load.
2. Low amplification factors, low plate resistance, and high plate current.
3. The curvature of the lower portion of the curve.
4. The load impedance is twice the plate resistance.
5. Because appreciable plate current flows during the entire grid-voltage cycle.
6. $e_p = E_b - i_p R_L$.
7. Zero.
8. 400 volts.
9. 30 ma.
10. 4 percent.
11. The primary-to-secondary turns ratio is equal to the square root of the primary-to-secondary matching impedances.
12. The loss in voltage through the transformer primary and secondary leakage reactance as a result of (1) load current and (2) capacitive current due to shunting capacitance.
13. High primary inductance and low leakage inductance.
14. (1) Second harmonics and all even-numbered harmonics are eliminated from output; (2) plate power supply hum is reduced; (3) no d-c core saturation is present; and (4) no regeneration is caused by signal currents in the power supply.
15. They are 180° out of phase.
16. About 72 percent.
17. Because the plate current flows for a smaller part of the input cycle.
18. Approximately 2.6 times as much.
19. Because the plate load resistance is twice as great.
20. 30 db.

21. Because the human ear responds logarithmically to changes in sound levels.
22. Zero reference level.
23. One watt.

CHAPTER 7

OSCILLATORS

1. The feedback must be regenerative, and the feedback energy must be sufficient to compensate for the energy losses in the grid circuit.
2. The distributed inductance and capacitance of the circuit components and the interelectrode capacitance of the tubes.
3. Measurement of the d-c voltage developed across the grid resistor.
4. By the mutual coupling between $L1$ and $L2$.
5. The flywheel effect (interchange of energy between the tank coil and capacitor).
6. It makes the oscillator self-starting.
7. The d-c component of plate current is isolated from the tuned circuit in the shunt-fed Hartley oscillator.
8. The Colpitts oscillator uses a split-tank capacitor instead of a split-tank inductor.
9. Through the plate-grid interelectrode capacitance of the tube.
10. The circuit having the higher Q.
11. In the high- and ultrahigh-frequency ranges.
12. Oscillator in L, $L2$, CT, control grid, and screen grid; amplifier in the plate, $C4$, and $L3$.
13. The frequency is increased.
14. By an increase in screen potential and an equal decrease in suppressor bias via $C1$, causing a decrease in screen current.
15. Quartz is more rugged and has a higher Q.
16. The X-cut crystal has a negative temperature coefficient (an increase in temperature causes a decrease in frequency) and the Y-cut crystal has a positive temperature coefficient (an increase in temperature causes an increase in frequency).
17. The AT-cut.
18. By placing the frequency-determining elements in a temperature-controlled oven.
19. Because the tank suddenly "looks like" a capacitor instead of an inductor and the feedback becomes negative instead of positive.
20. Because the required feedback is less.
21. The RC time constant of the circuit.
22. The frequency is increased.
23. The thyratron is more stable and the deionizing time is reduced
24. The linearity of the output voltage will be improved.

25. (See fig. 7–21.)
26. The values of R and C in the coupling networks.
27. To force them to operate at the synchronizing frequency.
28. In the cathode circuit or between grid and cathode.
29. The multivibrator will synchronize at a frequency that is some subdivision (for example ½) of the synchronizing frequency.
30. As electronic switches to produce gate voltages.

CHAPTER 8

MODULATION AND DEMODULATION

1. In the side bands.
2. 100 percent.
3. To prevent interference with other channels.
4. By the modulation transformer, M, in series with the tank and $B+$.
5. 2,000 volts.
6. The a-f input is equal to one-half the r-f input.
7. 4,000 volts.
8. 22.4 percent.
9. The output r-f current is reduced to zero.
10. It is four times as great.
11. Class C.
12. Space, weight, and input power are less than for plate modulation.
13. The degree of modulation, the power, and the intelligibility are reduced.
14. Because a buzzer or audio oscillator having a constant amplitude output voltage is used as the tone source.
15. The m-c-w receiver tuning is broader.
16. The amplitude of the modulating frequency.
17. It is taken from the carrier and redistributed in the side bands.
18. The approximate bandwidth is equal to the sum of the modulation frequency and the carrier frequency deviation.
19. Because the transmitter input to the antenna is constant and independent of the modulating signal.
20. To increase the initial frequency deviation of the oscillator to a suitable value in the output.
21. Its reactance varies with the modulating signal and thereby varies the frequency of the oscillator stage.

22. To ensure that only the amplitude, and not the instantaneous frequency of the modulating signal, will influence the extent of the carrier frequency swing.
23. One whose current-voltage relation is not a straight line.
24. Sum-and-difference frequencies (side-band frequencies) and a zero frequency or d-c component.
25. The waveform is distorted and new frequencies are produced.
26. Because the output voltage is essentially proportional to the square of the effective input voltage.
27. One volt.
28. Square law.
29. Contact potential.
30. Because it normally handles large input signals with minimum distortion.
31. To maintain the voltage across R during the time when no plate current flows.
32. The values of R and $C2$.
33. (1) It draws power from the preceding tuned circuit and therefore the circuit Q, the sensitivity, and the selectivity are reduced; (2) its interelectrode capacitance limits its usefulness at high carrier frequencies; (3) it distorts on weak signals; and (4) considerable amplification is required ahead of the detector.
34. A diode detector and a triode amplifier.
35. Because it draws grid current.
36. Because it does not draw grid current.
37. The nonlinearity of the response curve.
38. Because the discriminator is sensitive to both amplitude and frequency changes.
39. Upon the ratio between e_5 and e_6.

CHAPTER 9

TRANSMITTERS

1. Because c-w has fewer side bands and therefore greater signal strength in the remaining side-band frequencies.
2. Because the antennas are too long.
3. The signals are capable of being transmitted through magnetic storms that blank out the higher r-f channels.
4. The upper end of the ultrahigh-frequency band.

5. The large number of crystals needed.
6. By placing the frequency-determining components of the oscillator in a temperature-controlled oven, loading the oscillator very lightly, and isolating it with a buffer stage.
7. The electron-coupled oscillator (ECO).
8. So that the crystal may be operated at a lower frequency and therefore be larger and more rugged.
9. One-fourth.
10. Inversely.
11. (1) High-grid driving voltage, (2) high grid bias, and (3) the plate tank tuned to the desired harmonic.
12. Because it is supplied at a time when the plate voltage is (and hence plate losses are) at a minimum.
13. In the grid resistor.
14. It causes a loss in grid current and operating bias. Plate current then becomes dangerously high and the tube may be damaged.
15. They are larger and heavier.
16. It is prevented from breaking into oscillation.
17. Because there is additional power loss in the screen-grid circuits.
18. Parasitic oscillations.
19. By placing an inductor and resistor in parallel and inserting them in the grid and plate leads of an r-f amplifier of the transmitter.
20. Blocked-grid keying, as shown in figure 9–14, B.
21. So that plate voltage cannot be applied to *V*3 before the proper filament and bias voltages are developed.
22. High plate current, high plate dissipation, power loss, and low output.
23. Plate current cannot be brought to the proper minimum, and grid current will be too low or may even reverse.
24. (1) Greater range, (2) less interference, (3) smaller and much simpler to operate, and (4) more c-w transmitters than radio telephone transmitters may operate in a given band.
25. Low.
26. The range of frequencies over which the microphone is capable of responding should be no wider than the over-all frequency response of the system, and the response should be essentially uniform (free from sharp peaks or dips) throughout its range.
27. To prevent the transmission line capacitance from effectively shunting the microphone and causing an increased voltage loss, especially at the high-frequency end of the audio band.
28. 1 dyne per square centimeter input and 1 milliwatt output.

29. Less gain is required in the amplifiers with consequent greater margin over thermal noise, amplifier hum, and noise pick-up in the line between the microphone and the amplifier.
30. It requires an external voltage source, may be noisy, and may have mechanical resonance at certain frequencies.
31. The impedance is low.
32. Sensitivity to high temperature, humidity, and rough handling.
33. They are resistant to vibration, shock, and rough handling.
34. By the zero mutual inductive coupling between the plate tank inductor and the antenna coupling coil at audio frequencies.
35. In order to generate the required frequency deviation of ± 15 kc on either side of the carrier.

CHAPTER 10

TRANSMISSION LINES

1. They may be used as (1) impedance-matching devices, (2) phase shifters and inverters, (3) wave filters and chokes, and (4) oscillator frequency controls.
2. The characteristic impedance is increased.
3. They are in phase.
4. Because of line losses.
5. The characteristic impedance.
6. Maximum.
7. Minimum.
8. Maximum.
9. Zero.
10. They are equal.
11. Because of circuit losses.
12. A parallel-resonant circuit having high resistance.
13. An inductor.
14. A series resonant circuit.
15. The voltage and current curves remain the same as for the open-end line except that they are shifted toward the output end by an amount that increases as the capacitive reactance is reduced. (See fig. 10–7, A.)
16. The ratio of the effective voltage at a loop to the effective voltage at a node; (2) the ratio of the effective current at a loop to the effective current at a node; and (3) the ratio of the characteristic impedance to the impedance of the load.

17. The SWR is increased.
18. High radiation loss at high frequencies.
19. Rubber insulation losses.
20. The two conductors are balanced (same capacitance) to ground.
21. Higher.
22. The voltage level at which the waveguide arcs over.
23. The guides would be too large.
24. In the *TE* (transverse electric) mode of operation the *E* lines lie in transverse planes containing the *X* and *Y* axes, and the *E* lines also are parallel with the *Y* axis and perpendicular to the *Z* axis.
25. In the *TM* (transverse magnetic) mode of operation the magnetic field lies in transverse planes that contain the *X* and *Y* axes and that are wholly transverse to the guide axis.
26. The mode having the lowest cutoff frequency for a given size of guide is called the DOMINANT MODE.
27. Minimum.
28. A probe is inserted in a narrow slot that is parallel to the electric field, and a crystal or rectifier bolometer are used to detect the signal.
29. Lecher lines are used as tuned-circuit elements and as resonant lines for the purpose of obtaining wavelength.
30. The power-handling capacity is reduced, the efficiency is lowered, and the resistance is increased.
31. The impedance is high.
32. Because they are too long.
33. Because the impedance (the $\frac{E}{I}$ ratio) varies widely between the shorted and the open end.

CHAPTER 11

ANTENNAS AND PROPAGATION

1. A moving electric field creates a magnetic field and a moving magnetic field creates an electric field.
2. $\lambda = \frac{300}{f}$.
3. Sine waveform.
4. At few wavelengths.
5. Inversely.

6. Because they have large standing waves of voltage and current with a minimum of generator current and voltage.
7. Less.
8. The antenna input impedance.
9. Because the large-diameter radiator has greater capacitance hence less inductance.
10. Radiation resistance is the value of resistance that will dissipate the same power that the antenna dissipates.
11. (a) 73.2 ohms; (b) 36.6 ohms.
12. The ohmic resistance of the antenna conductor; corona discharge; and insulator losses.
13. They are parallel.
14. Horizontal.
15. It is one-half wavelength long, or any even or odd multiple thereof; it may be mounted either vertically or horizontally and need not be connected conductively to the ground.
16. It is grounded and is one quarter-wavelength long, or any odd multiple thereof.
17. By adding an inductor in series with the antenna.
18. By adding a capacitor in series with the antenna.
19. R.
20. Because the ground-reflected component (1) is shifted in phase 180° upon reflection, (2) has the same magnitude as the direct component, and (3) travels a path of approximately the same length as that of the direct component.
21. Because greater energy is absorbed from the wave at the higher frequencies.
22. Much higher.
23. The D layer.
24. (1) The angle at which the sky wave strikes the ionosphere; (2) the frequency of the transmission; and (3) the ion density.
25. It decreases.
26. It decreases the range to that of surface-wave ranges.
27. The F_2 layer.
28. No.
29. Warm layers of air are found above cooler layers.
30. It increases the range many miles beyond the normal range.

CHAPTER 12

ELEMENTARY COMMUNICATIONS RECEIVERS

1. Selectivity is the ability to select the desired signal; sensitivity is the ability to amplify weak signals.
2. Because, unlike triodes, they usually do not require neutralization, and they have higher gain.
3. Because of the high amplification through the multistage amplifier.
4. Decoupling circuits.
5. To spearate stations that are crowded together on the dial.
6. (1) To rectify the signal, and (2) to filter it (remove the r-f component and pass the a-f component on to the a-f amplifier).
7. The curved portion near the cutoff point.
8. The type of reproducer.
9. They are equal.
10. To the a-c signal current in the voice coil.
11. The selectivity is not constant over the tuning range.
12. It isolates the oscillator from the antenna-ground system, and reduces the strength of images.
13. An unwanted signal that always differs from the desired station by an amount equal to twice the intermediate frequency.
14. To provide a low impedance (coupling between the lower end of *L2* and the grounded end of *C2*), thus bypassing the decoupling filters in the a-v-c circuit.
15. By means of the electron stream itself.
16. To provide frequency stability for the local oscillator—that is, to prevent it from locking in synchronism with the station frequency signal.
17. Amplification and mixing.
18. The i-f output voltage to the r-f input voltage.
19. The class-A voltage amplifier.
20. To increase the receiver selectivity. The crystal has a higher Q than tuned circuits employing inductors and capacitors.
21. Because the diode detector has good linearity and can handle relatively large signals without overloading.
22. By reducing the time constant of R_1C_2, for example, by reducing *R1* to 100 k-ohms.
23. Variable-mu tubes.

24. From the output of the filter $R1$, $C2$, $C3$ connected across the diode load resistor $R2$.
25. (a) 3 volts; (b) cathode bias.
26. Ordinary a-g-c attenuates all signals, even weak ones. Delayed a-g-c does not attenuate weak signals—only strong ones.
27. When it acts as an open circuit.
28. To beat with the c-w signal to produce an audible tone.
29. To disable the receiver when no signals are being received.
30. The method of detection used.
31. To discriminate against images and to increase the amplitude of weak signals.
32. By using the second harmonic of the local oscillator for mixing, by connecting capacitors having a negative temperature coefficient across those having a positive temperature coefficient, and by using proper voltage regulation and choice of oscillator tubes.
33. Because the gain of each f-m stage is less than that of the corresponding a-m stage.
34. To provide for the discriminator a signal having constant amplitude.

CHAPTER 13

ELECTRONIC TEST EQUIPMENT

1. To produce a visual presentation of circuit waveforms.
2. By comparing the observed waveforms with the optimum efficiency waveforms printed on the schematic diagrams or on the equipment.
3. Electrostatic and electromagnetic.
4. Electromagnetic deflection.
5. A cylindrical grid surrounding the cathode.
6. To accelerate the electrons in the beam and aid in the focus action.
7. Because of their high velocity in the direction of the screen.
8. By varying the voltage between the first anode and the cathode.
9. By varying the current through the focus coil.
10. The angle through which the beam may be deflected in any direction from the center line through the tube.
11. The length of time that the screen glows at the point where the electron beam strikes it.
12. By applying the sine-wave voltage to the vertical deflection plates and at the same time applying a saw-tooth voltage of the same frequency to the horizontal deflection plates.

13. To prevent excessive attenuation by the vertical amplifier circuits.
14. To blank out the return trace of the sweep generator.
15. By the use of a voltage divider.
16. (1) The distance in millimeters that the spot will move on the screen when 1 volt is applied across one pair of deflection plates, and (2) the input voltage to the amplifier (horizontal or vertical) for a deflection of 1 inch of the spot on the screen.
17. So that the entire pulse will be visible on the screen, including the leading edge.
18. Observing (1) the waveforms of short pulses, (2) the time interval of the pulses as well as the time between pulses, and (3) the standing-wave ratio in a waveguide between a radar transmitter and antenna.
19. To cut an amplifier tube, *V*1, on and off at the multivibrator frequency, as in figure 13–9.
20. They tend to detune the self-excited oscillator circuits to which they are coupled.
21. Because energy is absorbed from the oscillator tank by the circuit under test thus decreasing the a-c component of plate voltage and decreasing the feedback energy so that the grid is driven less positive.
22. *WWV*.
23. The output is too low.
24. The alignment of r-f and i-f circuits; and, if sufficiently accurate, the determination of unknown frequencies by the zero-beat method.
25. To isolate the fixed and variable oscillators and to prevent them from pulling into step with each other as they approach the same frequency at low a-f settings on the dial.
26. A portable radio receiver and a meter to indicate the strength of the received signal.
27. To permit calibration of the field-intensity meter since the output of the shot-noise generator is known.
28. To prevent local oscillator radiation from the antenna circuit.
29. To provide a visual indication of the frequency components of an amplitude-modulated radio wave.
30. The voltage drops must be equal and they must be in phase.
31. Tests to determine the turns ratio of transformers, and capacitor quality.
32. The emission-type tester indicates the relative value of a tube in terms of cathode emission; the transconductance tester indicates not only this value but also the ability of the grid voltage to control the plate current.

33. A radio receiver.
34. An a-c signal applied to the grid.
35. Because the relatively low impedance of the voltmeter causes excessive current to flow through the circuit under test, thus causing an excessive voltage drop and an abnormal distribution of the circuit voltage.
36. The metallic rectifier capacitance shorts the high frequencies around the meter.
37. To introduce a contact potential across the grid of *V2B* to balance the bridge.

CHAPTER 14

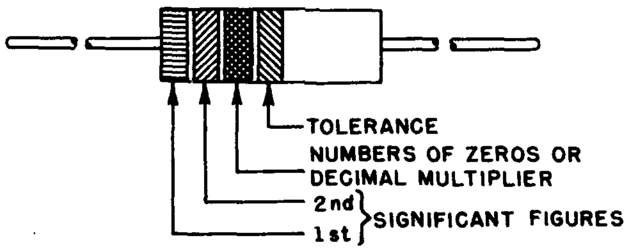
INTRODUCTION TO RADAR

1. 1,100 feet.
2. Direction, distance, and height.
3. Doppler effect.
4. The distance traveled by the signals.
5. Pulse radar does not depend on the relative frequency of the returned signal or on the motion of the target.
6. Search, fire control, and fighter-director.
7. To control and direct aircraft in air-to-air attack and defense; in some instances they are used as search radars.
8. Range.
9. Range and azimuth angle.
10. 328 yards.
11. 6.1 microseconds.
12. The time is the same in both cases.
13. The operator turns a handcrank to move the range indicator, or gate, to the target and then reads the range in yards from the counter assembly.
14. The accuracy is increased.
15. The altitude is equal to the slant range multiplied by the sine of the angle of elevation.
16. PPI.
17. (1) In the range scope it is amplified and applied to the vertical deflection plate in such a way as to produce a pip on the horizontal time base line on the screen; (2) in the PPI scope it is amplified and applied to the control grid in such a way that the trace is brightened momentarily on the radial time base line on the screen.

18. High power, wide beam angle, and long pulse widths.
19. Relatively low power, short pulse width, and narrow beam angle.
20. It produces the synchronizing signals that trigger the transmitter; triggers the indicator sweep; and coordinates the other associated circuits.
21. To generate r-f energy in short, powerful pulses.
22. The higher the carrier frequency, the smaller is the antenna array.
23. The power rating is reduced.
24. The ratio of the resting time to the pulse width.
25. The persistence of the screen and the rotational speed of the antenna.
26. The width of the transmitted pulse (receiver is blocked during transmitter pulsing periods).
27. The duty cycle.
28. A negative-going square wave of sufficient amplitude.
29. Upon the size of the cylinder, the strength of the magnetic field, and the difference in potential between the cathode and plate.
30. Because of the relatively long resting time between pulses.
31. To prevent the outgoing energy from entering the receiver and to prevent the energy of the echo from entering the transmitter.
32. Because of the higher frequencies used in radar.
33. By the use of many i-f stages.
34. (1) Repetition-rate oscillator, (2) saw-tooth generator, (3) multivibrator, (4) magnetron, and (5) local oscillator.
35. To reshape waveforms from a generating circuit according to the needs of the system.

APPENDIX II

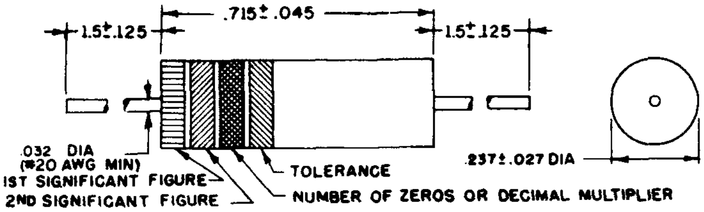
ELECTRONIC COLOR CODING AND SYMBOLS



COLOR	SIGNIFICANT FIGURE OR NUMBER OF ZEROS	DECIMAL MULTIPLIER	RESISTANCE TOLERANCE
			PERCENT ±
BLACK	0	---	---
BROWN	1	---	---
RED	2	---	---
ORANGE	3	---	---
YELLOW	4	---	---
GREEN	5	---	---
BLUE	6	---	---
VIOLET	7	---	---
GRAY	8	---	---
WHITE	9	---	---
GOLD	---	0.1	5(J) *
SILVER	---	---	10(K) *
NO COLOR...	---	---	20(M) *

* SYMBOL DESIGNATION ALTERNATE FOR COLOR

Figure A-1.—Standard resistor color code.



NOTES:

1. All dimensions in inches.
2. Referenced specification shall be of the issue in effect on date of invitation for bids.

TYPE-DESIGNATION EXAMPLE

RC30 — STYLE; BF — CHARACTERISTIC; 153 — RESISTANCE; M — RESISTANCE TOLERANCE

CHARACTERISTIC (MAXIMUM AMBIENT TEMPERATURE)

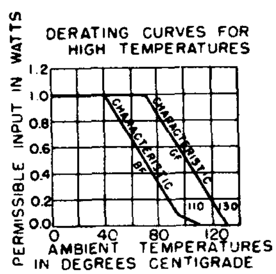
Symbol	Maximum ambient temperature for full-load operation
	° C.
B	40
G	70

RESISTANCE TOLERANCE

Symbol	Resistance tolerance
	Percent (±)
J	5
K	10
M	20

CHARACTERISTIC (RESISTANCE-TEMPERATURE)

Nominal resistance	Maximum allowable change in resistance from resistance at ambient temperature of 25° C. Symbol F: At -55° C. (ambient)	At 105° C. (ambient)
Ohms	Percent (±)	Percent (±)
1,000 and under	6.5	5
1,100 to 10,000	10	6
11,000 to 0.1 megohm	13	7.5
Megohms		
0.11 to 1.0	20	10
1.1 to 10	26	18
11 and over	35	22



Power rating	1 watt
Minimum resistance value	10 ohms 1/
Maximum resistance value	20 megohms 1/
Continuous working voltage (maximum) 2/	500 volts

1 FOR STANDARD RESISTANCE VALUES, SEE MS91374 IN MIL-R-11A.

2 CONTINUOUS WORKING VOLTAGE SHALL BE COMPUTED IN ACCORDANCE WITH THE FOLLOWING FORMULA BUT IN NO CASE SHALL IT EXCEED 500 VOLTS:

$$\text{VOLTAGE} = \sqrt{\text{POWER RATING} \times \text{NOMINAL RESISTANCE}}$$

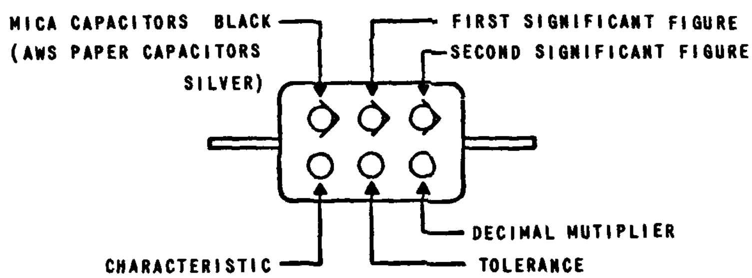
Figure A-2.—Specifications of composition resistors.

axial leads	color	radial leads
Band A	indicates first significant figure of resistance value in ohms.	Body A
Band B	indicates second significant figure.	End B
Band C	indicates decimal multiplier.	Band C or dot
Band D	if any, indicates tolerance in percent about nominal resistance value. If no color appears in this position, tolerance is 20%.	Band D

Note: Low-power insulated wire-wound resistors have axial leads and are color coded similar to the left-hand figure above except that band A is double width.

Courtesy Telecommunication Laboratories, Inc.

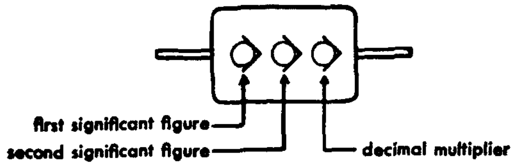
Figure A-3.—Color code for axial- and radial-lead fixed composition resistors.



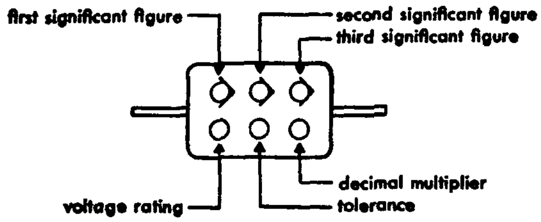
RMA 6-dot

Courtesy Telecommunication Laboratories, Inc.

Figure A-4.—AWS and NME color code for fixed mica capacitors.



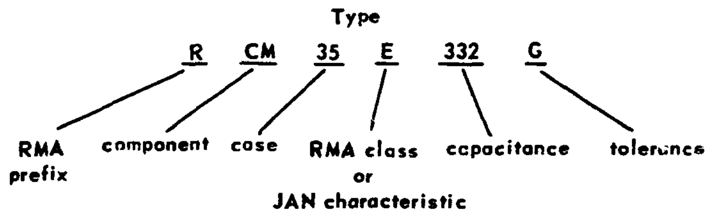
RMA 3-dot 500-volt, ± 20% tolerance only



RMA 6-dot

Examples

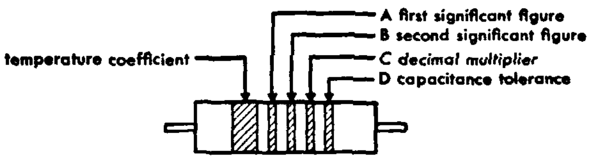
type	top row			bottom row			description
					tolerance	multiplier	
	left	center	right	left	center	right	
RMA (3 dot)	red	green	brown	none	none	none	250 $\mu\mu f \pm 20\%$, 500 volts
RMA	brown	black	black	blue	green	brown	1000 $\mu\mu f \pm 5\%$, 600 volts
RMA	brown	red	green	gold	red	brown	1250 $\mu\mu f \pm 2\%$, 1000 volts
CM30B681J	black	blue	gray	brown	gold	brown	680 $\mu\mu f \pm 5\%$, characteristic B
CM35E332G	black	orange	orange	yellow	red	red	3300 $\mu\mu f \pm 2\%$, characteristic E



Courtesy Telecommunication Laboratories, Inc.

Figure A-5.—RMA 3-dot and 6-dot color code for fixed mica capacitors.

color	significant figure	multiplier	capacitance tolerance in % c > 10 μμf	capacitance tolerance in μμf c ≦ 10 μμf	temperature coefficient parts/million/° C
black	0	1	±20	2.0	0
brown	1	10	±1		−30
red	2	100	±2		−80
orange	3	1,000			−150
yellow	4	—			−220
green	5	—	±5	0.5	−330
blue	6	—			−470
violet	7	—			−750
gray	8	0.01		0.25	+30
white	9	0.1	±10	1.0	−330 ±500



Examples

wide band	narrow bands or dots A	B	C	D	description
black	black	red	black	black	2.0 μμf ± 2 μμf, zero temp coeff
blue	red	red	black	green	22 μμf ±5%, −470 ppm/° C temp coeff
violet	gray	red	brown	silver	820 μμf ±10%, −750 ppm/° C temp coeff

Courtesy Telecommunication Laboratories, Inc.

Figure A-6.—Color code for fixed ceramic capacitors.

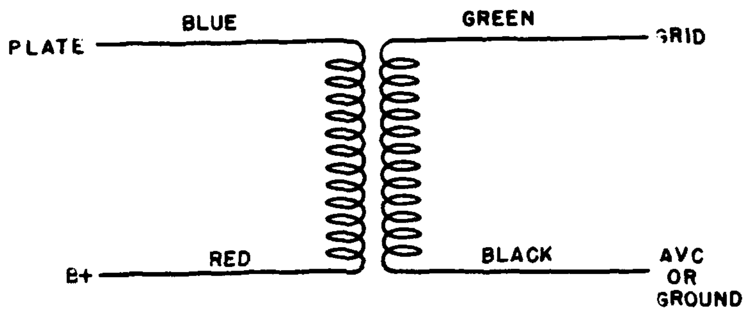
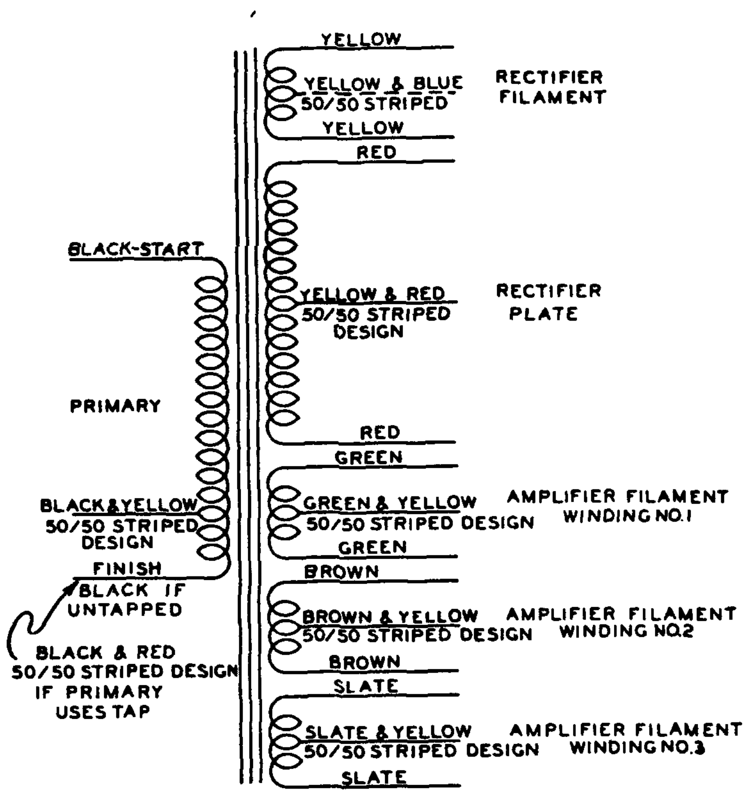
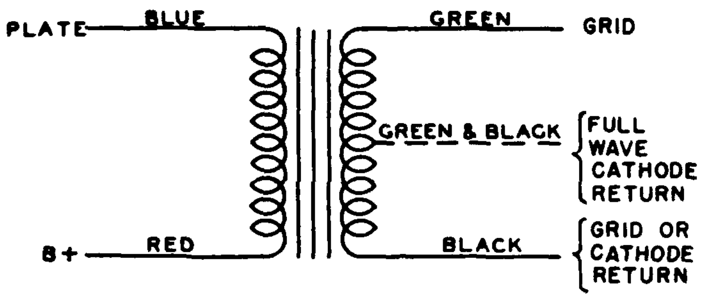


Figure A-7.—RMA color code for i-f transformers.



Courtesy P. R. Mallory & Co., Inc.

Figure A-8.—RMA color code for power transformers.



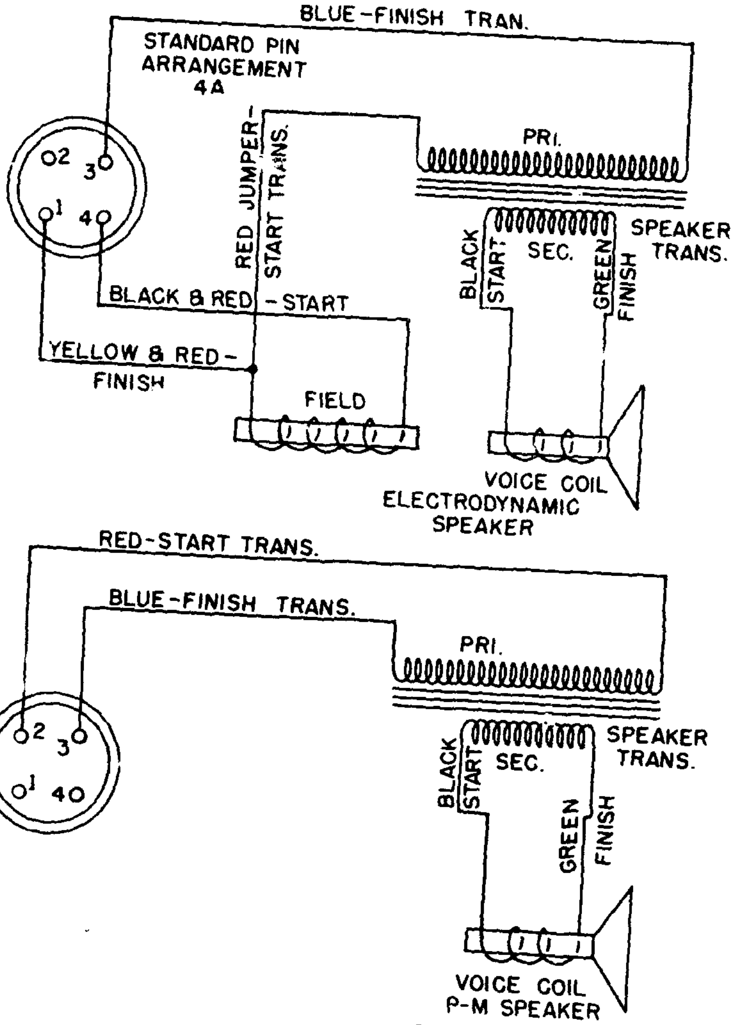
Courtesy P. R. Mallory & Co., Inc.

Figure A-9.—RMA color code for interstage audio transformers.

Standard colors used in chassis wiring for the purpose of circuit identification of the equipment are as follows:

Circuit	*Color*
Grounds, grounded elements, and returns	Black.
Heaters or filaments, off ground	Brown.
Power supply B plus	Red.
Screen grids	Orange.
Cathodes	Yellow.
Control grids	Green.
Plates	Blue.
Power supply, minus	Violet (purple).
A-c power lines	Gray.
Miscellaneous, above or below ground returns, a-v-c, etc.	White.

For other electrical and electronic symbols refer to Military Standard, MIL-STD-15A, 1 April 1954.



Courtesy P. R. Mallory & Co., Inc.

Figure A-10.—Standard pin arrangement and color code for electrodynamic and P. M. speakers.

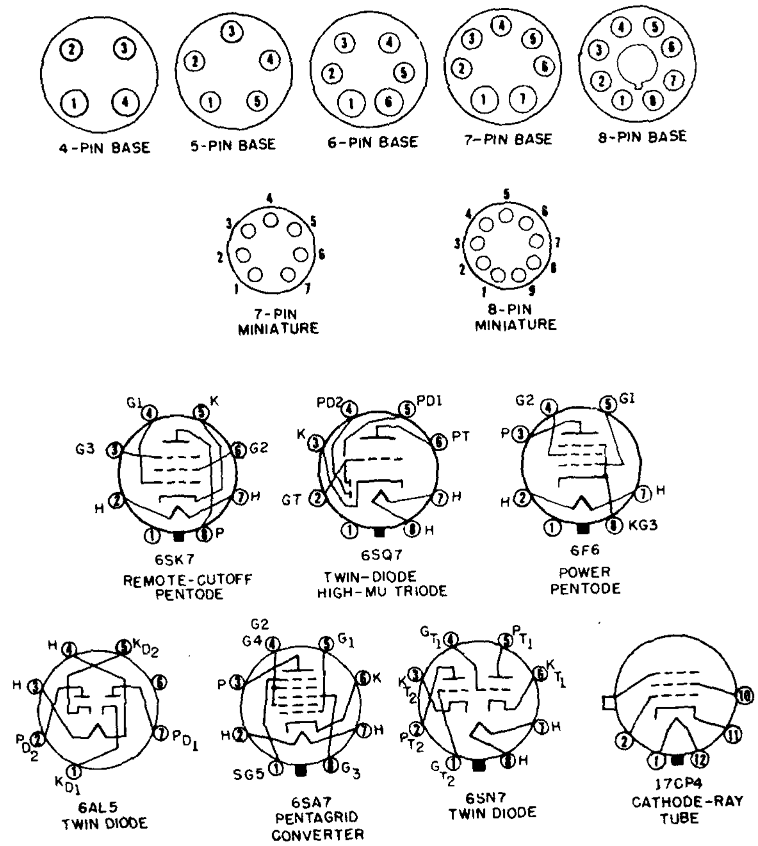


Figure A-11.—Common electron-tube bases showing arrangement of pins as viewed from the bottom.

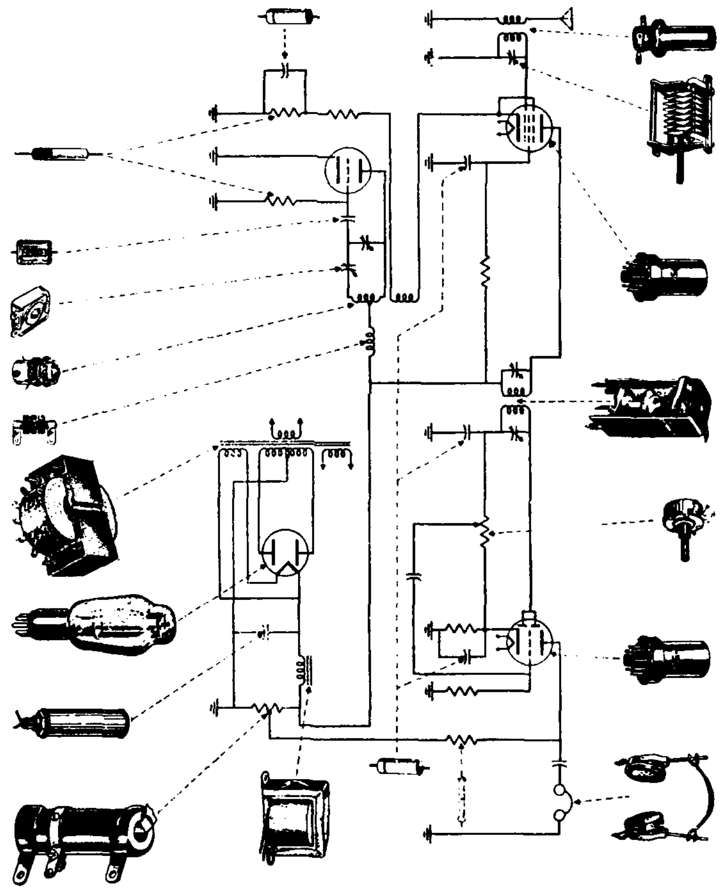


Figure A-12.—Commonly used circuit components and their corresponding symbols.

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